62 IRE 7.S1

IRE STANDARDS ON ELECTRON TUBES METHODS OF TESTING 1962

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* Approved by the IRE Standards Committee, May 11, 1961.

Preface

The word *standard*, as used in this document, means that the definition, method of measurement or test has been investigated carefully and endorsed by the IRE Standards Committee and the IRE Electron Tubes Committee. The methods of measurement are recommended practices, which, with reasonable care, will permit objective measurements of performance.

This "IRE Standards on Electron Tubes: Methods of Testing, 1962" supersedes "Standards on Electron Tubes: Methods of Testing, 1950, Parts I and II (50 IRE 7.S2)."

This Standard combines the material published in the above Standard together with many revisions and new headings, in addition to a number of definitions which do not appear in "IRE Standards on Electron Tubes: Definitions of Terms, 1957 (57 IRE 7.S2)."

The IRE Electron Tubes Committee would appreciate hearing from users of this document relative to criticisms and suggestions for future proposed methods of test.

GEORGE A. ESPERSEN, *Chairman*
IRE Electron Tubes Committee

Contents

Part 1

Conventional Receiving Tubes

Subcommittee 7.1

Tubes in Which Transit-Time is Not Essential

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1. INTRODUCTION

This Standard deals with the methods of measurement of the important characteristics of electron tubes. Terms used in this Standard are defined in the "IRE Standards on Electron Tubes: Definitions of Terms, 1957, 57 IRE 7.S2."[1]

1.1 General Precautions

Attention is called to the necessity, especially in tests of low-power apparatus, such as receiving tubes, of eliminating or correcting for errors due to the presence of the measuring instruments in the test circuit. This applies particularly to the currents taken by voltmeters and other shunt-connected apparatus, and to the voltage drops in ammeters and other series-connected apparatus.

Attention is also called to the desirability of keeping the test conditions, such as filament heating, plate potential, and plate current, within the safe limits specified by the manufacturers. If the specified safe limits are exceeded, the characteristics of the electron tube may be permanently altered and subsequent tests vitiated. When particular tests are required to extend somewhat beyond a specified safe limit (see Sections 2.1 and 3), such portions of the test should be made as rapidly as possible and preferably after the conclusion of the tests within the specified safe limit.

1.2 General Test Conditions

Except when the nature of a test calls for varying or abnormal conditions, all tests should be made at the normal rated conditions specified by the manufacturers of the electron tubes. If the manufacturer's rating is not specific, test conditions not specified should be selected in accordance with the best judgment of the tester and should be clearly and fully stated as a part of the test data. In general, measurements should be made after the tube has attained normal operating-temperature.

[1] PROC. IRE, vol. 45, pp. 983–1010; July, 1957.

When a filament is rated in both voltage and current, the rated voltage should be employed in tests. When filaments are to be used in series, the rated current may be employed, but this condition of measurement is to be stated as a part of the test data. Direct current should be used for filament heating, except where the normal operating condition is with alternating-current heating, in which case the use of the latter should be stated. When dc heating is employed, the negative filament terminal should be taken as the datum of potential. If the proper filament terminal to be used as the negative one is not indicated by the manufacturer or specified in any recognized standard manner for a given vacuum-tube structure, the terminal used as the negative one should be stated with the test data. When ac heating is employed for a filamentary cathode, the midpoint (*i.e.*, the center tap on the filament-transformer secondary, or the midpoint on a resistor shunting the filament) should be taken as the datum. It should be noted that these ac and dc heating potential datum conditions are not equivalent and should not be expected to give equivalent readings. If substantially equivalent readings are desired for the two cases, the datum of potential for ac heating must be taken at a point where the direct potential is more negative than that of the filament midpoint by an amount numerically equal to one half the root-mean-square value of the filament voltage. In the case of indirectly heated equipotential cathodes, the cathode is taken as the datum of potential. The connection of the cathode in any part of the heater circuit will usually have no effect upon the measured characteristics.

2. FILAMENT OR HEATER CHARACTERISTICS

2.1 Filament or Heater Electrical Characteristics

Readings of filament or heater current and voltage are taken with voltage applied only to the filament or heater terminals. Measurements should be made over a range of filament voltage or current from values too low to give appreciable electron emission in service to

at least the safe maximum voltage or current.[2] The current should be measured when it has reached equilibrium and should be corrected for the current drawn by the voltmeter. Curves should be plotted with values of filament voltage as abscissas and values of filament current and filament power as ordinates.

The resistance of a cold filament or heater is much smaller than its resistance under normal operating conditions. If the filament of a large tube is connected directly to the heating source, excessive filament current may flow and cause damage to the tube or test equipment. Therefore, it may be necessary to limit the filament current to some specified starting value.

2.2 Filament or Heater Heating Characteristic

When tubes are operated with the filaments or heaters in series, it is desirable that the voltage be divided during the heating period as nearly as possible in accordance with the rated operating voltages of the individual filaments or heaters, in order to minimize the likelihood of burning out and to insure minimum heating time of the entire complement. In order to judge the heating characteristic of individual tubes in a series string, the time variation of the filament or heater resistance of each tube should be compared with the average of the entire complement.

Measurements are made in a circuit such as that shown in Fig. 1, and the resulting data are plotted as per cent of rated or final heater resistance against time as in Fig. 2. In Fig. 1 T_1 and T_2 are a variable and a tapped transformer, respectively, having ratings adequate to insure good regulation. Switch S_1 is for energizing the complete circuit, and the combination of switch S_2 and resistor R_m is used to protect the meter A against excessive initial surges, the resistance R_m being made equal to the impedance of the meter A. Protection against short circuits is afforded by the fuse.

The output voltage of T_2 is to be equal to the rated heater voltage of the tube plus the drop through the low-resistance meter A or its equivalent R_m. Readings of current I are taken at known intervals. The percentage of the final resistance attained at the time of any reading will be 100 V/IR_f, where R_f is obtained from the final heater voltage at rated heater current.

The curves of Fig. 2 show the percentage of final heater resistance plotted against heating time. Curve A may be considered as average for indirectly heated tubes used in series operation, while curves B and C are for heaters having relatively fast and slow heating characteristics, respectively (when operated in series with tubes having heating characteristics such as A, Fig. 2). It will be noted that a tube such as that having the characteristic B (Fig. 2) will operate with heater temperature above the final value during part of the heating cycle, and the maximum voltage across the heater will exceed the rated value in proportion to the amount by which the curve of per cent of final heater resistance against time differs from the average of that characteristic for all tubes in the circuit. On the other hand, a tube having a characteristic like that of C (Fig. 2) will reach its final temperature more slowly when operated in a series circuit than the average tube, and the heater voltage will never exceed the rated proportional value.

2.2.1 Heater Warmup Time for Series-String Operation: A special test method is used for obtaining heater warmup time for tubes with indirectly heated cathodes for series operation. The method has been established empirically and provides a useful index of comparison for heater performance during warmup in series-string circuits. The circuit for measuring heater warmup time is shown in Fig. 3. The tube to be tested is placed in the circuit cold with switch S open and with E_s and R adjusted to the values shown. Switch S is then closed, and the time t_w is measured from the closing of the switch until the voltmeter V reaches the value V_1. This time t_w is defined as the heater warmup time of the tube.

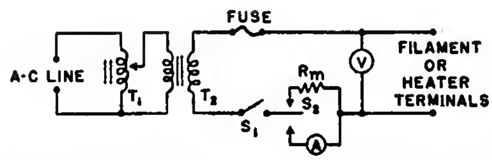

Fig. 1—Circuit arrangement for measuring filament or heater heating characteristic.

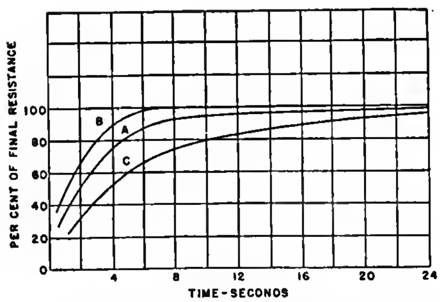

Fig. 2—Curves showing percentage of final heater resistance against heating time.

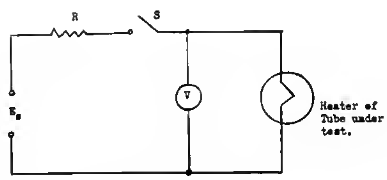

Fig. 3—Circuit for measuring heater warmup time. E_s = applied voltage rms or dc value; R = total series resistance including supply; E_f = rated heater voltage; I_f = rated heater current; V_1 = heater test voltage = 5.0 $E_f/6.3$; E_s = 25 $E_f/6.3$; R = 3.0 E_f/I_f.

2.3 Cathode Heating Time

The cathode heating time is arbitrarily taken as the time required for the time rate of change of the cathode current to reach maximum value. All applied voltages are to remain constant during the measurement. The

[2] See Section 1.1, General Precautions.

electrodes must be at room temperature before the test is made.

A sample plot of cathode current aginst time is given in Fig. 4. From this it is seen that the current increases slowly at first, rises increasingly rapidly, and then gradually reaches its final value. The maximum time rate of change of cathode current referred to above corresponds to the point of maximum slope or point of inflection of the curve of Fig. 4. Measurement under this definition may be made by either of two circuits which give comparable results. Method A is sometimes used but the more recent Method B is recommended for new equipment, as it is free of the possibility of difference in saturation effects between different transformers.

2.3.1 Method A: The instantaneous current flowing in the secondary of the step-down transformer of Fig. 5 depends only on the rate of change of the current in the primary and is independent of its final value. Hence, the time of the maximum rate of change will be indicated by the maximum deflection of the meter needle. The speed at which the meter needle moves is indicative of the acceleration and has no bearing on the problem; only the time required for the needle to reach the peak of the swing is of importance.

The characteristics of the output transformer and/or meter should be specified. The meter should have a short period, and the step-down transformer should be selected to give convenient deflections.

2.3.2 Method B: In the circuit arrangement shown in Fig. 6, the meter preferably has a resistance of less than 1000 ohms and the low-leakage capacitor has a capacitance of about 8 μf. The shunting resistor is made variable to keep the indicator of the meter on scale. It is important that the meter used have good damping characteristics. The time is taken from the instant the filament or heater circuit is closed to the instant the capacitor charging current reaches a maximum. The circuit constants and the characteristics of the meter should be specified.

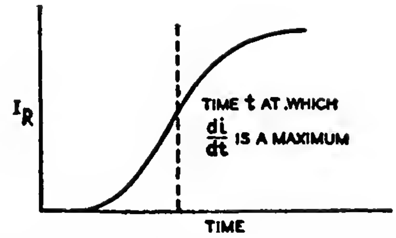

Fig. 4—Relation between cathode current and time.

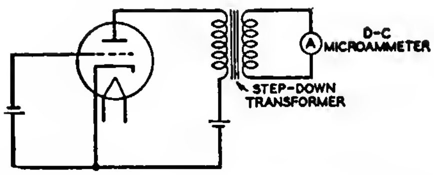

Fig. 5—Circuit arrangement for measuring cathode heating time (Method A).

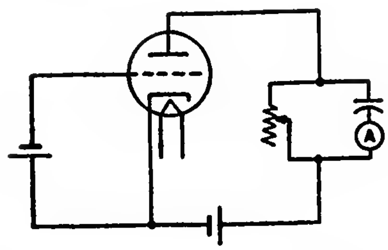

Fig. 6—Circuit arrangement for measuring cathode heating time (Method B).

2.4 Cathode Cooling Time

The cooling time can be taken as the time required for the time rate of change in cathode current to reach a maximum after the filament or heater circuit is opened. The circuit given in Fig. 5 is applicable. Alternatively, it may be taken as the time required for the current to reach half its steady value after the heater circuit is opened.

2.5 Operation Time

Operation time in a vacuum tube is defined as the time after simultaneous application of all electrode voltages for an electrode current to reach a stated fraction of its final value. In practice, the final value is taken as the value reached after a specified time interval long enough for a normal tube of the type under test to have reached a substantially stable value. All electrode voltages are to remain constant during the test. The tube elements should be at room temperature at the start of the test.

3. Emission Tests

The emission of a thermionic cathode may be evaluated by measuring the diode characteristic of the tube.

Two regions of the diode characteristic are generally of importance. One is the temperature-limited-emission region in which the emission is limited principally by the temperature of the cathode (region *A* in Fig. 7). The other region of interest is that in which the departure from the law of space-charge-limited current becomes noticeable (region *B* in Fig. 7).

These two regions may be represented quantitatively by two specific emission currents. Region *A* may be represented by the flection-point emission current shown as point *a* in Fig. 7. Region *B* may be represented by the inflection-point emission current shown as point *b* in Fig. 7.

3.1 Measurement of Flection-Point Emission Current

The flection-point emission current, sometimes used as an approximate measure of total emission or temperature-limited emission, is defined as the current at the point on the diode characteristic where the second derivative has its maximum negative value. The value of this current may be obtained with reasonable accuracy from a diode characteristic taken by a suitable

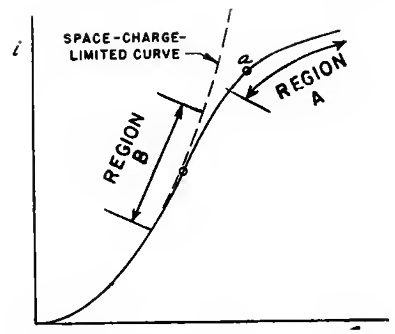

Fig. 7—Typical diode characteristic.

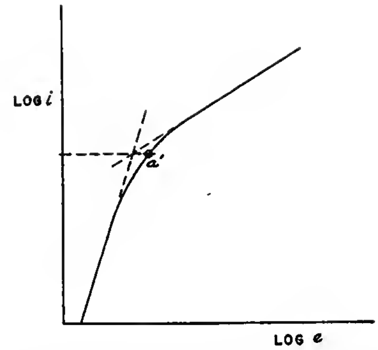

Fig. 8—Determination of flection-point emission current, slope-intersection method.

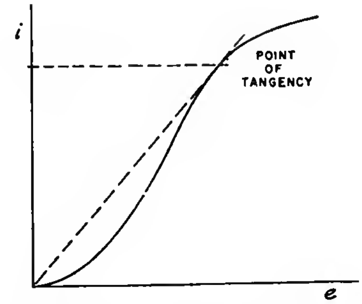

Fig. 9—Determination of flection-point emission current, tangent method.

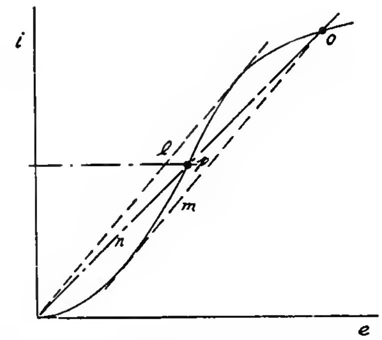

Fig. 10—Determination of inflection-point emission current, graphical method.

method, such as one of those described in Section 4.4.

3.1.1 Graphical Methods: An approximation of the flection-point emission current may be obtained from the diode characteristic by two graphical methods.

3.1.2 Slope-Intersection Method: The flection-point emission current is approximately equal to the current corresponding to the intersection of two straight lines representing the slopes of the diode characteristic in the space-charge-limited region and in the temperature-limited region, respectively (Fig. 8). This method is particularly effective if the diode characteristic is plotted on log-log paper. It tends to give a current value higher than the actual flection-point current.

3.1.3 Tangent Method: The point of tangency of a line drawn through the origin, tangent to the diode characteristic, as shown in Fig. 9, will indicate the approximate flection-point emission current. If a curve tracer is used to obain the diode characteristic, a rotatable line drawn on transparent material may be affixed to the screen of the indicator at the origin of the trace so that it can be manually adjusted to tangency with the trace. This method tends to give a current value somewhat lower than that corresponding to the flection point.

3.2 Measurement of Inflection-Point Emission Current

The inflection-point emission current is defined as the current at the point on the diode characteristic at which the second derivative is zero. The value of this current may be obtained with reasonable accuracy from a diode characteristic taken by a suitable method, such as one of those described in Section 4.4.

3.2.1 Graphical Method: An approximation of the inflection-point emission current may be obtained from the diode characteristic by a graphical method. The complete diode characteristic of the tube must be obtained and plotted. Then, as in Section 3.1.3, a line l is drawn through the origin, tangent to the characteristic, as shown in Fig. 10. Another line m is then drawn parallel to l and tangent to the lower part of the characteristic. From the point O, where m intersects the characteristic, a line n is drawn through the origin. The intersection p of line n with the characteristic will approximate the inflection point.

3.2.2 Break-Away Point: The inflection-point emission current is of particular interest for small-signal tubes used as linear amplifiers. Such tubes usually employ oxide-coated cathodes, in which the measure of inflection-point emission current is close to the so-called break-away point.

3.2.3 Determination of Break-Away Point: It can be shown experimentally that the break-away from the 3/2-power-law space-charge line in a diode characteristic of a tube with an oxide-coated cathode, plotted on two-thirds-power paper or on log-log paper, occurs very near the actual inflection point. Therefore, the break-away current i in Fig. 11 can be used as a measure of inflection-point emission current.

3.3 Comparison of Emission Currents of Tubes

When it is desired to compare the emission currents of several tubes of the same type at the flection point,

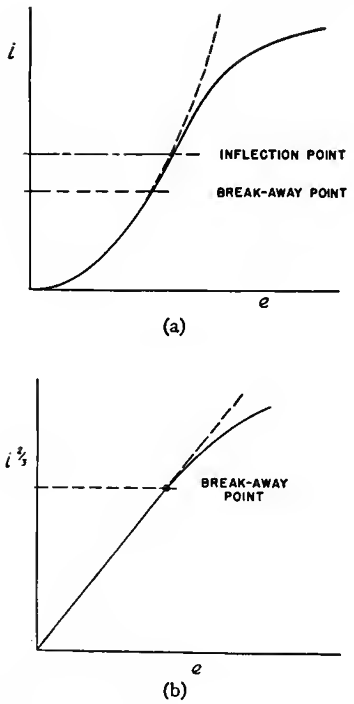

Fig. 11—Determination of emission break-away point.

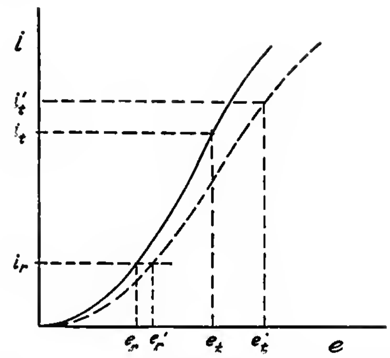

Fig. 12—Comparison of the emission currents of electron tubes.

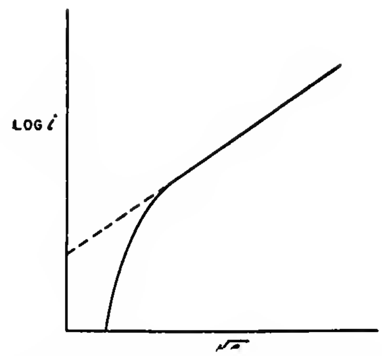

Fig. 13—Determination of field-free emission current.

inflection point, or any other specified point on the diode characteristic, such as point (e_t, i_t) in Fig. 12, it is possible to use the following procedure, in which the effect of small variations in cathode area and electrode spacings is eliminated.

A point (e_r, i_r) is selected on the space-charge-limited portion of the reference diode characteristic. To compare any other tube with the reference tube, it is first necessary to determine the voltage e_r' required to obtain the same current i_r. The proper voltage e_t' for the tube under test will then be given by the relation $e_t' = e_t(e_r'/e_r)$. The current i_t' will be a measure of the emission of this tube compared with the current i_t of the reference tube.

3.3.1 Precautions: The voltage e_r should be chosen sufficiently high so that contact potential is negligible in comparison. If this it not possible, the contact potential of each tube should be measured and the voltage e_r corrected for this effect.

The relatively large current obtainable in the inflection point region may cause a permanent change in the emission characteristic if the current is allowed to flow continuously. Since such a change must be avoided, pulse methods, described in Section 4.4, are desirable, and often essential.

3.4 Measurement of Field-Free Current

The value of field-free emission current of a cathode may be obtained from the diode characteristic. For this purpose the data are plotted as the logarithm of the current against the square root of the applied voltage; sufficient data beyond the flection point must be obtained to determine a straight line in this portion of the plot (Fig. 13). If this straight line is then extended to the point where it intersects the current axis, the value of current corresponding to this point will be the field-free emission current of the cathode. It may be found that sparking of the cathode, particularly with oxide-coated cathodes, occurs in the high-voltage region. Pulse methods, such as those described in Section 4.4, are ordinarily used for obtaining the data.

3.5 Emission Checks

For the purpose of making quick checks on the electron emission of a tube where apparatus for taking a complete diode characteristic is not available, or where a rough check is sufficient, one of the following methods may be used. It is necessary that the values obtained in such checks be correlated with results of more accurate tests in order for such information to be of value.

3.5.1 Direct Emission Check: For routine test purposes, electron emission may be checked with the filament or heater voltage adjusted to the rated operating value. Then all electrodes except the cathode are connected together to form a composite anode. A voltage, specified for the tube type under test, is applied to the composite anode, and a measurement is made of the electron current. The voltage may be direct or alternating.

In practice, anode potentials are chosen considerably lower than would be required for total emission. The choice of anode potential is determined to some extent by the sum of the maximum peak electrode currents

that will be required in service, or by the maximum voltage that can be applied for a reasonable length of time to tubes of a particular type without injury.

3.5.2 Indirect Emission Check: In tubes having filamentary cathodes that might be injured by passing a relatively large average emission current through the filament, it is frequently desirable to obtain an indirect check of the emission by noting the value of filament voltage for which a specified value of emission current is obtained. Provided a low value of anode voltage is employed, this method also avoids false readings caused by such effects as local overheating of the filament, gas currents, or field-emission currents.

3.5.3 Oscillation Emission Checks: The following method of checking emission, in which the tube is operated under specified conditions as a self-excited oscillator, is particularly adapted to use with power tubes. The filament voltage is reduced until the total radio-frequency power output has been reduced by a specified percentage, and the filament voltage is then measured. This value of filament voltage is an indirect measure of the filament activity. Alternatively, the filament voltage may be lowered to a stated fraction of its rated value and the decrease in output noted. These methods are arbitrary and give relative check results that are valuable only as long as individual tubes of the same type are compared. The results depend in a great degree upon the circuit conditions.

4. Characteristics of an Electron Tube

The static characteristics of an electron tube are valuable as a means of predicting the performance of the tube, since load characteristics may usually be computed for assumed circuit conditions. The various tube constants and the perveance may be calculated from families of static characteristics. Load characteristics may be obtained in some cases by actually operating the tube in the desired circuit and making the appropriate measurements.

4.1 Static Characteristics

The more useful static characteristics of an electron tube are: electrode characteristics, constant-current characteristics, and transfer characteristics.

In general, these characteristics may be obtained by dc methods up to the point where electrode dissipations exceed safe values. Pulse methods such as those described in Section 4.4 are necessary to obtain the information beyond this point. It is desirable to obtain static characteristics up to and slightly beyond the extreme conditions of voltage and current that the tube will experience in operation.

4.1.1 Direct-Current Method: A representative arrangement for the determination of the characteristics of vacuum tubes is shown in Fig. 14.

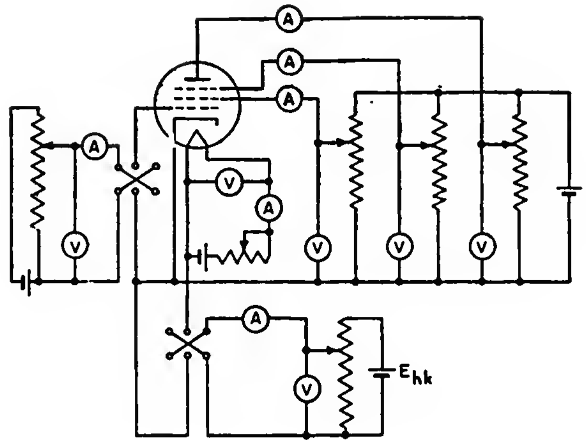

Fig. 14—Circuit arrangement for measuring static characteristics.

4.2 Load (Dynamic) Characteristics

4.2.1 Calculation of Load Characteristics from Static Characteristic Charts: Various forms of static characteristic charts can be conveniently used for precalculation or verification of tube performance by plotting load characteristics on them. In many cases the shape of the load characteristic can be predicted from the operating conditions and frequently is either a straight line or a portion of an ellipse. For example, class-A audio-frequency operation with resistive load may be represented as a straight line on a plate-characteristic chart. All classes of radio-frequency operation with resonant circuits in both input and output circuits are conveniently represented on a constant-current chart. Transfer characteristic charts are useful for class-B radio-frequency and audio-frequency plots when it is desired to examine the load characteristic directly for harmonics.

At frequencies where electron-transit-time effects become appreciable, it becomes inaccurate to calculate load characteristics from the static characteristics.

Load characteristics permit calculation of the entire performance data of the tube, such as input power, output power, efficiency, dissipation, excitation power, etc.

4.2.2 Direct Measurement of Load Characteristics: The load characteristics of a tube can be measured directly, without resort to calculation from the static characteristics. The tube should be set up for the required operating condition and the desired load characteristic observed by means of a cathode-ray oscillograph, the electrode voltage being applied to one pair of deflection plates and the voltage across a current-measuring resistor simultaneously applied to the other pair.

At frequencies at which electron transit time, tube capacitance, and tube lead inductance become important, they may have considerable effect upon the load characteristic. It is therefore advisable to take load characteristics at the frequency at which the tube is to be used.

4.2.2.1 Precaution: Care must be taken to insure that the measuring instrument and its connecting leads do not affect the shape of the load characteristic.

4.3 Perveance

4.3.1 Perveance of a Diode: The perveance of a diode with a unipotential cathode is given approximately by

$$G \simeq \frac{i_b}{e_b^{3/2}},$$

or more accurately by

$$G = 1/(3/2r_p)^{3/2} i_b^{1/2},$$

where i_b is the space-charge-limited current, e_b the anode voltage, and r_p the anode resistance.

The perveance, although ideally a constant which depends on the physical dimensions of the tube, actually varies somewhat with anode voltage and cathode temperature. Consequently, its value is of use primarily under specific conditions of cathode temperature and anode voltage. The numerical value of perveance of a diode can be calculated from measurements of voltage and current within the space-charge-limited region. If perveances at anode voltages on the order of one volt are desired, corrections for the effects of contact potential and initial electron velocities should be considered.

4.3.1.1 Graphical method: When current is plotted against the three-halves power of voltage, the slope of the curve is G. The slope is not constant over a wide range in voltage; consequently the perveance is determined at a desired value of voltage.

4.3.2 Perveance of a Triode or Multigrid Tube: The perveance of a negative-grid triode is given approximately by

$$G \simeq \frac{i_b}{e'^{3/2}},$$

where e' is the composite controlling voltage:

$$e' = \frac{\mu e_c + e_b}{\mu + 1}.$$

For multigrid tubes it is usually of sufficient accuracy to measure values of μ and the electrode voltages with the electrodes connected as a triode.

4.3.2.1 Low-voltage correction: When the diode voltage or composite controlling voltage has a low value (of the order of one volt), more accurate results are obtained by applying a correction for the effects of initial electron velocity and of contact potential difference. These effects are not readily separable, but their combined effect may be represented by a single voltage term ϵ. The composite controlling electrode voltage in this case is

$$e' = \frac{\mu e_c + e_b}{\mu + 1} + \epsilon,$$

and the perveance is

$$G = \frac{i_b}{\left[\dfrac{\mu e_c + e_b}{\mu + 1} + \epsilon\right]^{3/2}}.$$

Taking the transconductance g_m as i_b/e_c, and neglecting the partial derivatives of G, μ, and ϵ with respect to e_c, the correction voltage ϵ becomes

$$\epsilon \simeq \frac{(3/2)i_b\mu}{g_m(\mu + 1)} - \frac{\mu e_c + e_b}{\mu + 1}$$

$$G \simeq (1/i_b)^{1/2}\left[(2/3)g_m \frac{(\mu + 1)}{\mu}\right]^{3/2}.$$

When measured values of the terms on the right-hand side of the preceding expressions are used, the values of ϵ and G that may be computed are useful in problems of design and in the standardization of tube characteristics.

For multigrid tubes electrodes beyond the first grid are connected together, and the tube is measured as a triode.

4.4 Pulse Methods

It is of considerable importance to know the static characteristics of electron tubes in the region where electrode dissipation may exceed safe values. It is impossible to obtain these characteristics by the conventional direct-current methods without damaging the tube. In such cases it is necessary to employ pulse methods in which the tube is allowed to pass current only for short intervals of such duration and frequency that it is not damaged.

Pulse methods may be employed for obtaining electrode characteristics, transfer characteristics, or constant-current characteristics. The basic elements needed for a pulse method are the pulse generator and the current and voltage indicators. Where a single pulse generator is employed, it is usually connected to the control-grid circuit, as in Fig. 15, with the grid biased past cutoff in the absence of the pulse. If more than one pulse generator is used, as in Fig. 16, it is necessary to synchronize the pulses. In general, a single pulse generator is adequate and the single-generator method is simpler than the two-generator method.

Ordinarily, two methods are employed in obtaining characteristics, namely: the point-by-point method and the curve-tracer method. The point-by-point method consists of applying known pulse voltages to electrodes and simultaneously measuring the corresponding maximum pulse currents to the electrodes. The applied pulse voltages are adjusted point by point over the desired ranges to obtain a family of characteristic curves. The curve-tracer method consists of applying a pulse voltage of such magnitude that the entire range of the desired characteristic is covered and of such shape that the oscillograph indicator shows graphically the true relation between electrode current and voltage.

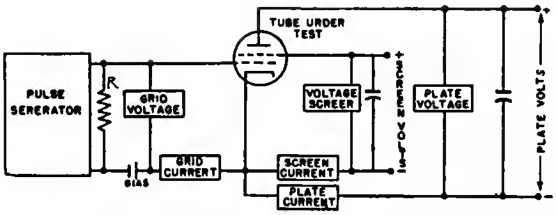

Fig. 15—Circuit arrangement for pulse measurement of tube characteristics, single-generator method.

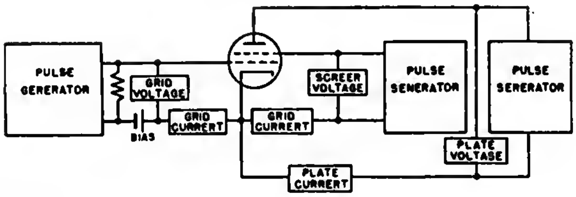

Fig. 16—Circuit arrangement for pulse measurement of tube characteristics, multiple-generator method.

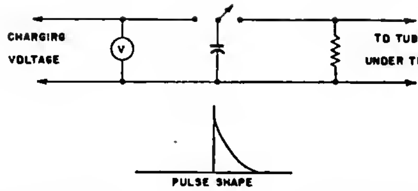

Fig. 17—Basic circuit arrangement for capacitor-discharge pulse generator.

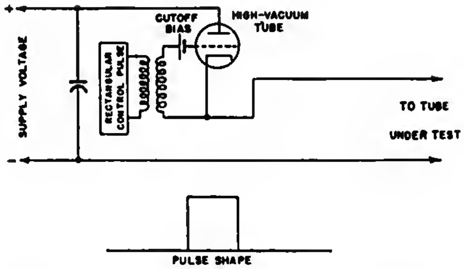

Fig. 18—Basic circuit arrangement for high-vacuum-tube pulse generator.

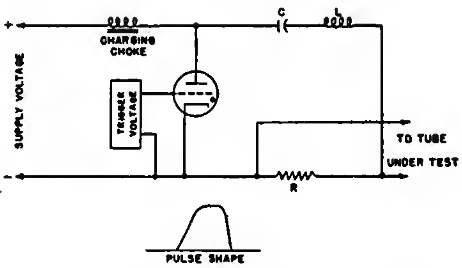

Fig. 19—Basic circuit arrangement for gas-tube *LC* pulse generator.

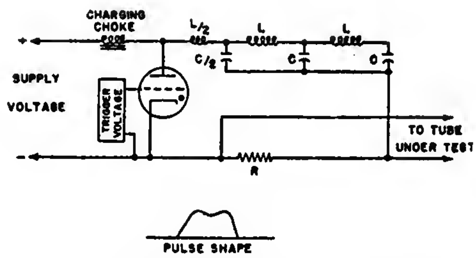

Fig. 20—Basic circuit arrangement for gas-tube pulse generator with pulse-forming line.

4.4.1 Point-by-Point Pulse Methods:

4.4.1.1 Types of pulse generators:

4.4.1.1.1 Capacitor-discharge type: The discharge of a capacitor provides one of the simplest means for obtaining a voltage pulse. The capacitor may be discharged either directly through the tube under test, or through a coupling device. In the absence of series impedance, the peak voltage applied to the tube is the voltage of the charged capacitor. Some means must be provided for switching the capacitor between charge and discharge circuits. The basic circuit is shown in Fig. 17. The switching device may be a mechanical switch, or an electron tube.

Because of the very short duration of the pulse voltage in a simple capacitor discharge, it is difficult to provide an accurate current indicator of good accuracy. It may therefore be desirable to shape the applied pulse so as to extend the duration of its peak. This may be done by the use of suitable networks and switching means. Figs. 18–20 show circuits for obtaining various pulse shapes. In the circuits of Figs. 19 and 20 the resistance of the parallel combinations of the coupling resistor R and the load of the tube under test should approximate the characteristic impedance $Z_0 = \sqrt{L/C}$ of the network. In all of the foregoing circuits, except that of Fig. 17, the pulse voltage must be measured directly across the tube under test by means of a suitable indicator.

In the circuit of Fig. 18 the vacuum tube may be considered as a pulse amplifier in which any pulse shape applied to the control-grid circuit is amplified and applied to the tube under test. This circuit has the advantage of not requiring a specific terminating impedance. Several amplifier tubes may operate in parallel to provide additional current if precautions usual for parallel operations are observed. There are several arrangements of this basic circuit that may be used, depending on the location of the ground connection and the polarity of the output pulse.

If secondary-emission effects in the tube under test introduce a negative impedance, the pulse generator must be shunted with a noninductive load resistor of such value as to maintain a net positive impedance at its terminals. This resistor R should be located between the generator and the current indicator, as shown in Fig. 15.

4.4.1.1.2 Sine-wave type: Synchronous mechanical contactors or properly controlled thyratrons in conjunction with alternating voltages can be used to apply a half-cycle pulse, or a smaller portion of a cycle, to the tube under test, as shown in Fig. 21. It is usually desirable to use only one out of every two or more cycles in

order to keep the electrode dissipation low. The system should have low internal impedance.

4.4.1.2 Types of indicators: Indicators are required for both current and voltage. In general, indicators for pulse methods may be divided into three types: dc meters, peak-reading meters, and oscillographs.

4.4.1.2.1 Direct-current meter indicators: Meters are the simplest indicators but have limited application for obtaining electron-tube characteristics by pulse methods. A dc voltmeter, may, for example, be used to read the capacitor voltage in the circuit of Fig. 17, a circuit that is sometimes useful in obtaining volt-ampere characteristics by point-by-point methods. The capacitor must have sufficient capacitance to supply the required charge, and the meter reading must be corrected for the voltage drop across any series impedance. Another circuit in which a dc meter gives satisfactory results is that of Fig. 23.

The dc meter may also be a useful type of indicator in the circuit of Fig. 15, where it may be used to indicate the voltage of the plate or screen. If high accuracy is desired, correction must be made for the voltage drop in the plate- or screen-current indicator.

4.4.1.2.2 Peak-reading voltage indicator: Peak-reading voltage indicators are useful in measuring the peak value of the pulse voltage applied to an electrode in obtaining vacuum-tube characteristics by point-by-point pulse methods.

In order to function successfully, the circuit consisting of the capacitor, resistance, and meter in Fig. 22 must have a time constant that is large with respect to the time interval between successive pulses. The indicator accuracy is greatest with rectangular pulses shown in Fig. 18, although good accuracy can be obtained with the pulse shapes shown in Figs. 19–21 if the pulse duration is sufficient.

Where the pulse is substantially square and the OFF/ON ratio is known, a moving coil or other averaging instrument may be used to measure the electrode current. The peak current is equal to the mean current (as measured by the meter) multiplied by the ratio of the total time to the pulse time.

The peak reading voltage indicator of Fig. 22 may be calibrated from a known dc source if correction is made for voltage drop in the high-vacuum diode. The usual practices of correcting for meter current or for voltage drop in current-measuring instruments should be followed. This type of indicator may be used also to determine peak current by measuring the peak voltage drop across a noninductive shunt resistor of known value.

The use of this indicator for current-measuring purposes may result in serious errors if the electrode characteristic of the tube under test exhibits a change of slope from positive to negative. Electrodes giving high-secondary-emission currents may have such characteristics, and a knowledge of the tube being tested or use of a cathode-ray oscillograph for checking purposes is desirable.

4.4.1.2.3 Cathode-ray-oscillograph current indicator: The indicator shown in Fig. 23 is useful in measuring the peak value of electrode currents in taking vacuum-tube characteristics by pulse methods based upon the circuits of Figs. 15 and 16. With large tubes, connection may in most instances be made directly to the vertical deflection plates of the oscillograph. Safety and good performance require that this method be applied only to those oscillographs in which connection may be made directly to deflection plates operated at or near ground potential.

It is usually desirable to utilize a linear horizontal sweep voltage, synchronized to the pulse-generator frequency, to spread out the current trace. In this manner the detection of possible errors caused by the negative slope of electrode characteristics is simplified. In

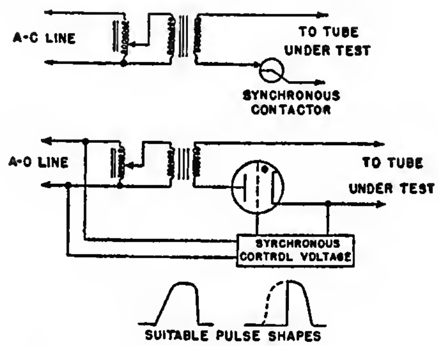

Fig. 21—Basic circuit arrangement for sine-wave-type pulse generator.

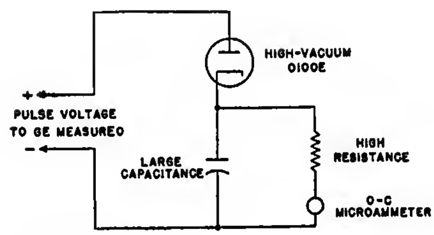

Fig. 22—Peak-reading voltage indicator.

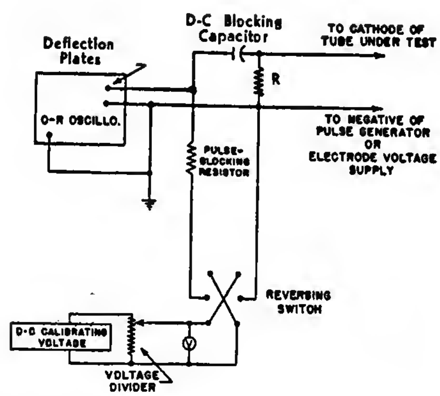

Fig. 23—Cathode-ray-oscillograph current indicator.

obtaining characteristics having only a positive slope, however, the peak current value may be obtained without a sweep voltage simply by measuring the height of the vertical deflection line. The dc calibrating circuit shown in Fig. 23 provides a convenient method of reading voltage drop produced in the resistor R by the electrode current. The use of a noninductive resistor is recommended, and connecting leads should be kept as short as possible. By means of the adjustable calibrating voltage, the cathode-ray beam may be depressed until the peak deflection coincides with the zero axis. The dc meter reading is then equal to the peak voltage drop in the current-measuring resistor if the reactance of the blocking capacitor is sufficiently small.

The same current indicator may be used to measure several electrode currents by switching it across various electrode-current-measuring resistors.

4.4.1.2.4 Cathode-ray-oscillograph voltage and current indicators: Fig. 24 shows the diagram of a type of indicator that is useful in taking vacuum-tube characteristics by either point-by-point or curve-tracer methods.

(Note: In point-by-point methods, the vertical deflection corresponding to the maximum horizontal deflection must always be considered in plotting characteristic curves. For characteristics having only positive slopes, this vertical deflection will also be the maximum.)

A complete trace may be obtained on the cathode-ray screen by applying a suitable voltage pulse to the grid or plate of the tube under test. In making measurements on large tubes, connections are usually made directly to the vertical and horizontal deflection plates. Amplifiers may be used if necessary, provided that they are designed with sufficient bandwidth and sufficiently linear phase-frequency response for the high-frequency components of the pulse.

For best results, the voltage attenuator should be capacitively balanced, as shown in Fig. 24. Noninductive resistors should be used and all leads kept as short as possible. Switches make possible the selection of the desired tube electrodes. The common method of calibrating the oscillograph screen by the use of known voltages to produce vertical and horizontal deflections may be used. Balancing of the voltage attenuator is achieved by making the product of resistance and capacitance of all sections alike. It is also important to take into consideration the input capacitance of the oscillograph. All leads must be properly shielded against stray fields.

The most reliable functioning of the indicator shown in Fig. 24 will be obtained with the common deflection-plate connection grounded. Since the voltage-attenuator current may then flow through the current-measuring resistors, suitable corrections must be made if the current is appreciable.

4.4.1.2.5 Cathode-ray-oscillograph voltage and current indicators with calibrating trace: The indicator shown in Fig. 25 may also be used in the same manner as that shown in Fig. 24, but means must be provided for measuring the deflection-voltage coordinates of any point on a trace. Two sources of independently variable direct voltage are required. Two calibrating voltages are applied simultaneously to the respective deflecting plates by synchronous switches. The calibrating voltages are timed to occur during the intervals between pulses. A calibrating spot will be obtained which can be shifted horizontally and vertically until it coincides with a desired point on the trace. The voltage coordinates of the point then correspond to the calibrating-meter readings. A correction is required for any drop that may occur in the synchronous switches. The calibrating voltage source must have sufficiently low impedance so that no appreciable change in calibrating voltage occurs during operation of the switches. The same precautions and considerations must be observed with this type of indicator as were described for the circuit of Fig. 24.

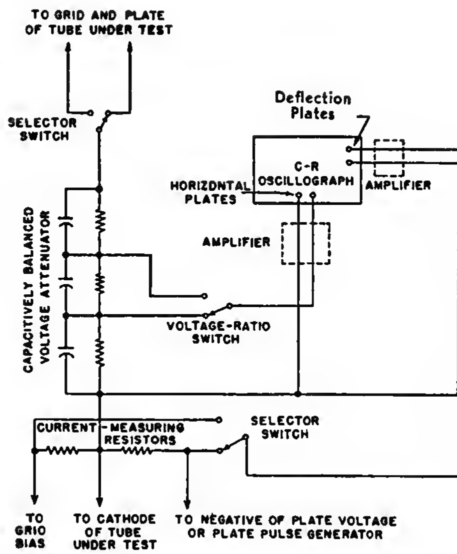

Fig. 24—Cathode-ray-oscillograph voltage and current indicator.

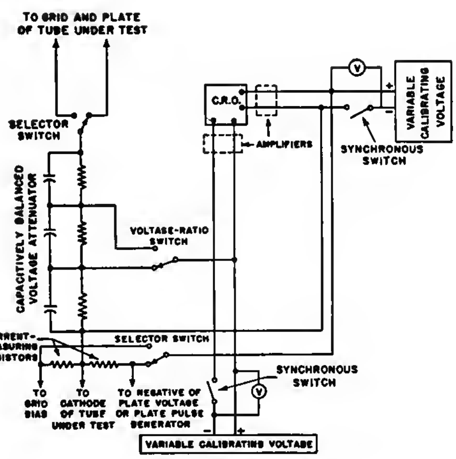

Fig. 25—Cathode-ray-oscillograph voltage and current indicator with calibrating voltages.

4.4.2 Curve-Tracer Method: It is sometimes of considerable advantage to view the complete characteristic of a tube in its useful range for the purposes of checking for irregularities, quickly checking performance, or recording characteristic data. For these purposes, the curve tracer affords a very convenient means of presenting the characteristic. An oscillographic current and voltage indicator is required for a curve tracer, and in most practical cases a cathode-ray-oscillographic indicator will be used. Where a permanent record is required, the screen of the cathode-ray tube may be photographed or the data may be transcribed from the screen point by point. Point-by-point transcribing of data from the tracer may be faster than making separate point-by-point measurements by other methods, and has the advantage that the whole trace may be viewed for irregularities that might escape another method.

The type of pulse generator employed for the curve tracer may depend upon the type of characteristic that is desired to view. In general, the pulse generator used for the independent variable should provide a triangular or sinusoidal pulse in order that the cathode-ray beam will be deflected at a nearly uniform time rate and thus provide uniform illumination of the curve. If pulse generators are used for other parameters, they should usually provide rectangular pulses of good regulation. Pulses of less than 10-μsec duration may result in loops in the trace because of phase effects in the input circuit of the tube under test. If it is necessary to use such short pulses, the circuits should be thoroughly checked for these effects.

4.4.2.1 Curve tracer for diode characteristics: A curve tracer suitable for viewing diode characteristics consists of a pulse generator capable of supplying the required peak voltage and current, and a cathode-ray-oscillograph current indicator such as described in Sections 4.4.1.2.4 and 4.4.1.2.5. Any of the pulse generators described in Section 4.4.1.1 having the desired wave shape may be used. When very high currents and voltages are required, a circuit similar to that of Fig. 18 involving one or more amplifier tubes driven by a sinusoidal or triangular pulse may be advantageous.

4.4.2.2 Curve tracer for electrode characteristics: In viewing electrode characteristics on a curve tracer, the sinusoidal or triangular pulse voltage is applied to the electrode under test. All other electrode voltages are held constant at the desired values during this pulse. It is necessary that the control grid be biased so that all current is cut off when no pulse is applied to the electrode under test. When the electrode under test is not the control grid, the control grid must be brought to the desired voltage and held at this value throughout the pulse. The regulation requirements for this voltage are severe. Other electrodes may be supplied from dc sources having large capacitances across them. If it is desired to view a family of curves, the parametric voltage must be changed in steps for successive pulses. This may be accomplished by means of synchronous contactors, or by means of electron tubes. If mechanical methods are used, precautions must be taken to prevent sparking of contacts.

4.4.2.3 Curve tracer for transfer characteristics: A satisfactory curve tracer for viewing conventional transfer characteristics is relatively easy to provide, since the pulse is applied to the control grid of the tube under test and the pulse generator need supply only the relatively low peak control-grid current. The plate and other grids, if any, may be supplied from dc sources shunted by large capacitors in order to maintain the voltage constant. If it is desired to view a family of curves, the parametric voltages may be varied by synchronous means, as described in the previous section. Transfer characteristics from electrodes other than the control grid may be obtained by a method similar to that described in Section 4.4.2.2.

Since a curve tracer for transfer characteristics is simpler to build and operate than one for other types of characteristics, and since data obtained from transfer characteristics may be replotted to show electrode characteristics or constant-current characteristics, all necessary information can be obtained most easily in this manner.

4.4.2.4 Curve tracer for constant-current characteristics: Although it is possible to construct a curve tracer for showing constant-current characteristics, it is simpler to plot such curves from transfer characteristics obtained by the methods described in the previous section.

5. Residual Gas and Insulation Tests

Vacuum tubes of the same type having satisfactory static and cathode characteristics may differ from one another in the degree of vacuum and of insulation between tube elements. These properties may affect the tube life and the relative freedom from complications in tube operation. The degree of vacuum can be estimated by measuring the gas (ionization) current; the insulation can be judged from the leakage current in the tube. In many cases, it is convenient and sufficient to measure both currents in the circuit of the negatively biased control grid.

To separate the gas current and leakage current it is necessary to vary the operating voltages. Variation of operating voltages, however, may cause the temperature of parts of the tube to vary. Because both gas current and leakage currents are dependent on temperature to an extent determined by the design of the tube, it is impossible to separate a grid current into its various components with any degree of accuracy.

5.1 Total Current to a Negatively Biased Control Grid

A minute electric current can always be observed in the circuit of the negatively biased control grid of a vacuum tube. Usually it consists of several component currents arising from various physical phenomena. The currents are superimposed upon one another and do

not all flow in the same direction. These currents may result from the following causes: 1) electrons from the cathode that reach the grid by virtue of contact potentials and initial velocities; 2) gas (ionization); 3) leakage; and 4) primary electron emission from the control grid. Currents arising from these causes are shown in Fig. 26.

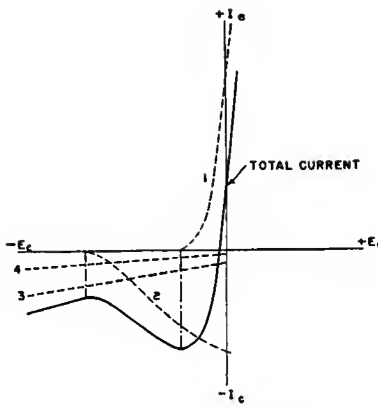

Fig. 26—Total grid current and its components. Curve 1—Electrons from the cathode that reach the grid by virtue of contact potentials and initial velocities. Curve 2—Ionization current. Curve 3—Leakage current. Curve 4—Electron emission from the control grid.

5.1.1 Measurement of Total Negative Grid Current:

5.1.1.1 Direct method: With a negative bias on the control grid, voltages are applied to other electrodes to establish a suitable electron current. After thermal equilibrium is reached, the current in the grid circuit is measured. One circuit arrangement is shown in Fig. 27.

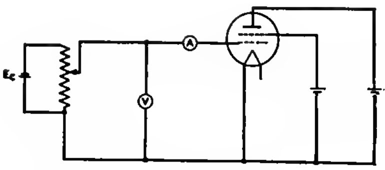

Fig. 27—Circuit arrangement for measuring total negative grid current, direct method.

5.1.1.2 Indirect method: A sensitive method of measuring total control-grid current, which is especially useful when the current is too small for convenient direct measurement by ordinary deflection instruments, is illustrated in Fig. 28. With the switch S closed and the grid and plate voltages adjusted to the desired values, the plate current is read. R_c is then inserted into the grid circuit by opening the switch, and the grid bias E_c is readjusted so that the plate current returns to its former value. The grid current can be computed from the change in grid voltage ΔE_c necessary to maintain constant plate current, since

$$I_c = \Delta E_c / R_c.$$

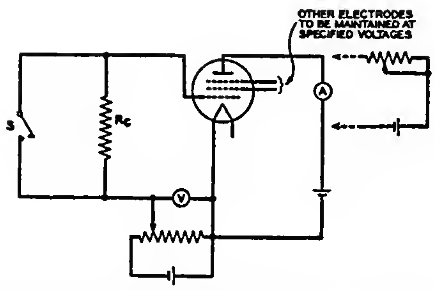

Fig. 28—Circuit arrangement for measuring small control-grid currents, indirect method.

The necessary value of R_c will depend upon the current to be measured. When a number of tubes of the same type are to be compared for grid current, it is often sufficient to estimate the relative grid current by noting the change in plate current when S is opened or closed.

The sensitivity of the method can be greatly improved by balancing out the normal plate current by the arrangement shown by the dotted lines of Fig. 28, in order to permit the employment of a more sensitive plate-current meter.

The leakage resistance across the switch S should be large in comparison with R_c.

5.2 Measurement of Gas (Ionization) Current

When the electrode voltages and currents are high enough, direct measurements under the desired operating conditions may be made with a microammeter, as explained in Section 5.2.1. In many cases there will not be sufficient ionization for these measurements at rated voltages. To check gas currents it is then necessary to use an indirect method such as those described in Sections 5.2.2 and 5.2.3. These methods serve to compare relative ionization in tubes of a given type, but do not afford a direct comparison of gas currents under operating voltages.

5.2.1 Subtraction Method: Gas current can be determined by the circuit of Fig. 27. First, the total negative grid current is measured as described above; grid bias is then increased to, or slightly beyond, the cutoff point, and the grid current is again noted. If the primary grid emission is relatively small, gas current is approximately equal to the difference between the two readings. This is so because the gas current is included in the first reading, whereas there is no gas current during the second measurement as there is no electron current to produce gas ions. The second reading includes leakage current and primary grid-emission current. Leakage current may be assumed to be proportional to the grid voltage.

See 5.1.1

(*Note:* If the primary grid emission is not small, the results obtained by this method may be misleading. The primary grid-emission current will increase with grid voltage and may cause the difference between the two readings to be zero or even negative. Other methods are then required to evaluate the results.)

5.2.2 Ionization-Gauge Method: In many types of

vacuum tubes, the degree of vacuum can be estimated without measuring the negative grid current. The method is based upon a circuit exemplified by Fig. 29, in which a positive voltage greater than the ionization potential is applied to the number-one grid and a negative voltage is applied to one or more other electrodes. Positive ions formed by the collision of electrons with gas molecules are collected by the negative electrodes. The resulting positive-ion current gives an indication of the gas density, and therefore gives a means for comparison of the degree of vacuum of individual tubes of the same design. Since the nature of the gas affects the measurement, it must be considered when measurements are made with different gases.

5.2.2.1 Precaution: Although the cathode temperature in this test must be adjusted to give a sufficiently high grid current for convenient readings, the grid current should not be too high because bombardment of the grid may cause gas evolution. The degree of vacuum in tubes of the same design may be compared if the grid voltage is kept constant and if constant grid current is maintained by adjusting the cathode temperature. Retarding-field oscillations that may occur during measurements in this circuit must be avoided by proper choice of voltage.

5.2.3 Dynamic Method: Gas current to the negatively biased anode can be measured in the circuit of Fig. 32, described in connection with the measurement of primary grid emission. In the application of this circuit to gas-current measurement, the anode circuit (not shown in Fig. 32) must include a suitable source of negative direct potential and a sensitive dc meter for reading the ion current. An advantage inherent in this method is that gas current can be measured as a function of grid input power. Furthermore, by plotting a graph of gas current vs grid electron current, one may determine the input power at which evolution of gas from the grid just begins. At this value of power, the graph begins to depart markedly from a straight line and bends upward (Fig. 30).

(*Note:* When a negative anode potential is used in the measurement of gas by methods 5.2.2 and 5.2.3, photoelectric emission from the anode may obscure gas-current reading. If photoemission is taking place, it may be recognized by projecting visible light from an external source onto the anode. If no change in gas current occurs under fixed potentials, photoemission may be assumed to be negligible. Moreover, if the grid power is sufficient to heat the anode to the point of thermionic emission, the electron current from the anode may affect the gas-current reading. This condition may be recognized by the reversal of anode current when the anode potential is gradually reduced to zero.)

5.3 Leakage Currents

Leakage currents should be measured with the insulating materials of the tube at or near operating temperatures. A suitable potential is applied between each two electrodes in turn and the current measured with the other electrodes floating.

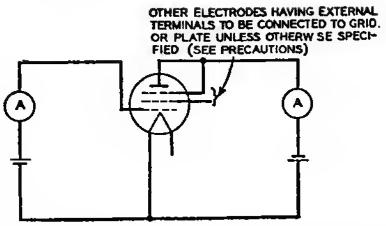

Fig. 29—Circuit arrangement for measuring gas current, ionization-gauge method.

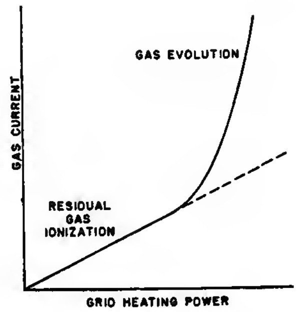

Fig. 30—Gas current as a function of grid dissipation.

Since these measurements may include emission currents from various electrodes, care should be taken that conditions are such as to preclude such emission. These conditions may be approached by heating the tube under normal operating conditions, turning off the filament or heater, and making measurements as soon thereafter as the electrodes have cooled to the point where electron-emission currents will be negligible.

5.3.1 Grid Leakage: Grid leakage may be estimated by testing in the circuit of Fig. 27 and applying the following general rules: Leakage currents are generally but not invariably proportional to the voltage applied. In small tubes where the grid-to-cathode gap is small, grid emission will generally depend on cathode temperature to a much greater extent than leakage or gas current. At any given anode or screen voltages, the gas current varies with the space current in the region of the electrode concerned; in triodes or pentodes it is proportional to the cathode current.

5.3.2 Heater-Cathode Insulation: Leakage current between the heater and the cathode of a tube should be measured with the rated heater voltage applied (sufficient time being allowed for the heater and cathode to attain normal temperature. The heater voltage may be either alternating or direct, as convenient). A voltage is applied in series with a microammeter between the heater and the cathode and the current is measured (due allowance being made for any voltage drop across the protective resistor). Measurements of heater-cathode leakage current may be made with alternating voltage or with direct voltage of either polarity. Because the current obtained will rarely be the same under the three conditions, the conditions of test must be

specified. Similarly, where the heater voltage is appreciable compared with the measuring voltage, the point of connection to the heater circuit must be specified. The voltage on elements other than heater and cathode may influence the results because of stray electron currents, and should be specified in the test conditions.

6. Inverse Electrode Currents

Thermionic electron-emission currents from elements other than the cathode may be of sufficient magnitude to affect the performance of vacuum tubes. Test methods are described below for measuring thermionic emission from the control-grid, screen-grid, and anode of amplifiers and from the anode of diode rectifiers.

Secondary electron emission from control grids may also affect performance and control tests may be desirable.

6.1 Thermionic Grid Emission

During tube operation, the control grid is subject to heating by radiation and by electron bombardment when the grid potential becomes positive. As a result, the grid temperature may reach the level at which the grid begins to emit electrons thermionically. Consequently, an electron current from the grid to other electrodes may flow whenever the grid is negative with respect to these electrodes. The amount of such thermionic emission and the conditions under which its effect becomes appreciable can be determined by one of the following methods.

6.1.1 Subtraction Method: The total negative grid current measured beyond the cutoff point, as described under Section 5.2.1, consists of the leakage current and the thermionic grid-emission current. Since grid-leakage current can be measured independently (Section 5.3), it can be subtracted from the total measured current to give the value of thermionic emission current. If the leakage current is negligible, the original measurement gives the grid-emission current directly.

(*Note:* The grid-emission and leakage currents are both related to temperature. Therefore, the total grid current must be measured as quickly as possible after the cathode current is cut off. The value of leakage current measured must be corrected to the voltage at which the total current was measured.)

6.1.2 Direct Method: The connections for direct measurements of thermionic grid-emission current are shown in Fig. 31. During this test, the electrodes should be at their normal operating temperatures. To this end, it is recommended that the tube be operated in the usual way at its normal voltages for a time sufficient to insure normal temperature conditions. By means of switches, the circuit shown in Fig. 31 is then quickly established, and the grid emission is noted while the electrodes are still approximately at their normal temperatures. The cathode should be at its normal operating temperature throughout this test.

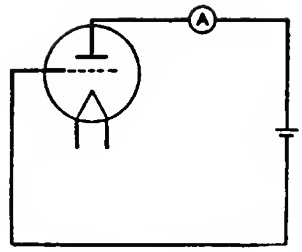

Fig. 31—Circuit arrangement for measuring grid-emission current, direct method.

6.1.2.1 Precautions: The above test gives the grid emission directly only when the leakage between the grid and plate electrodes is negligible. If the leakage current is not negligible but is known, it can be subtracted from the observed current to obtain the grid-emission current.

6.1.3 Dynamic Method: Measurement of thermionic grid emission as a function of grid dissipation power may be accomplished by use of the circuit shown in Fig. 32. In this arrangement the grid of the tube under test is heated by 60-cycle power from Transformer 1 on every alternate half-cycle through rectifier T_h and ammeter A_1. On the reverse half-cycle of Transformer 1, a sensitive meter A_2 in series with rectifier T_e indicates the thermionic grid-emission current (and leakage current if present). The battery B and the potentiometer R are included to provide a means of bucking out spurious currents around the T_h-T_e loop caused by contact potential and initial velocities. Voltmeter V in series with rectifier Tv serves to indicate the voltage on the grid on the heating half-cycle. With proper regard to the form factors of the various voltages and currents, the thermionic grid-emission current vs grid dissipation may be calculated. If a wattmeter is used in place of ammeter A_1 and voltmeter V, the grid dissipation can be read directly.

If necessary, grid emission can be measured in the tube with the anode heated to its normal operating temperature. A suitable circuit for this test is shown in Fig. 33. In this circuit a switching device, mechanical or electrical, is inserted into the grid circuit. In one position the heating current flows to the grid from the grid transformer through rectifier T_g and ammeter A_2 during one half-cycle of the 60-cycle supply. On the other half-cycle, the anode is heated from the plate transformer through rectifier T_p and ammeter A_3. After the electrodes have attained operating temperatures, the switch is thrown to its other position in which the grid is connected through sensitive meter A_1 to a negative voltage sufficient to cut off the plate current. The thermionic grid-emission current (including leakage current if present) under these conditions may be read on meter A_1.

6.2 Secondary Grid Emission

Secondary electron emission from the grid surface as a result of electron bombardment is an ever-present

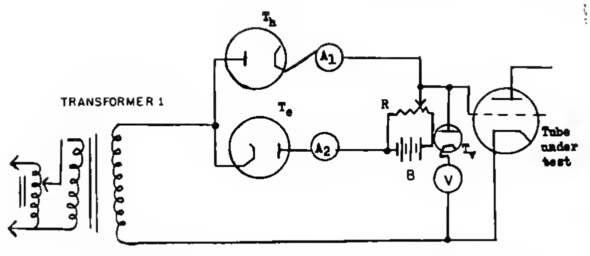

Fig. 32—Circuit arrangmement for measuring primary grid emission, dynamic method.

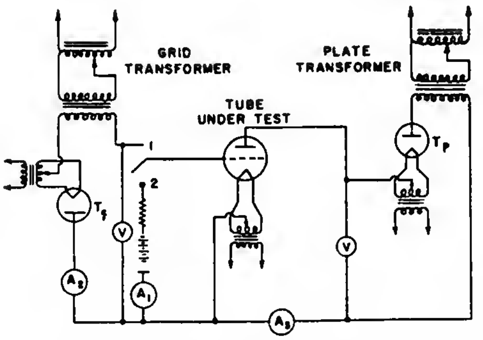

Fig. 33—Circuit arrangement for measuring primary grid emission with plate dissipation.

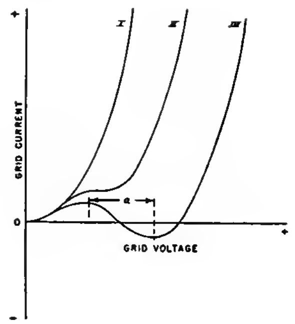

Fig. 34—Typical grid characteristics showing effects of secondary emission. Curve I—No appreciable secondary emission. Curve II—Appreciable secondary emission. Curve III—Pronounced secondary emission (note negative resistance in region a).

phenomenon in electron tubes when positive potentials are applied to the grid statically or dynamically. Secondary-electron-emission current flows from the grid to other more positive electrodes as long as electron current flows to the grid. In the external grid circuit the bombarding current and the secondary-emission current flow in opposite directions, and only their difference can be observed on a meter.

Pronounced secondary grid emission in a vacuum-tube amplifier or oscillator can help tube operation by increasing the output and efficiency or by reducing the grid-excitation power. It may, however, interfere with tube operation by preventing the realization of linear amplification or by causing parasitic oscillations. Secondary grid emission may exert a detrimental effect when it is of such magnitude as to produce a negative grid resistance. Negative grid resistance is evidenced by a negative slope of the curve of grid current vs grid voltage, or even by reversal of grid current. In Fig. 34 the effect of secondary grid emission on the grid characteristic is shown. Region a of curve III is a negative-resistance region.

The amount of secondary emission from the grid of a tube cannot be measured directly. Tubes of the same type, however, may be compared as to the relative amount of secondary grid emission present by projecting a grid characteristic upon the screen of a cathode-ray tube, as described in Section 4.4.2, or by checking by static means a point on the grid characteristic for chosen values of electrode voltages. If a static test is used, electrode voltages must be chosen that do not result in excessive electrode dissipation.

6.2.1 Calculation Method: The amount of secondary grid emission at any point on the tube characteristic may be calculated if the distribution of the thermionic electron current to the electrodes is known. The distribution of the thermionic electron current can be calculated approximately for various combinations of electrode voltages by use of empirical formulas. Knowledge of electrode spacings and dimensions is required for this calculation. This method, while not precise, is the only one available at the present time for determining the amount of secondary grid emission.

6.3 Reverse Emission in Rectifier Diodes (Back Emission)

Reverse emission in a rectifier is the inverse thermionic current between anode and cathode. If the dissipation during the conducting part of the cycle is sufficient to cause heating of the anode to a temperature high enough to enable it to emit thermionically, the released electrons will bombard the cathode during the time in which the cathode is positive with respect to the anode. A basic circuit suitable for the measurement of reverse emission is shown in Fig. 35. T_1, the tube under test, is operated as a half-wave rectifier and loaded by R_1 and C. T_h is a diode of impedance very low compared with that of T_1 and can conveniently be a gas-filled tube. The impedance of the rectifier T_e may be higher than that of T_h. R_2 is a limiting resistor to protect the meter A_2 in the event of flashover in T_1. The battery and the potentiometer R_3 are included to eliminate circulating currents in the T_h-T_e loop due to contact potentials.

During the conducting part of the cycle, T_1 and T_h conduct and, since T_h has a very low impedance compared with T_1, the time constants of the circuits will not be disturbed. During the nonconducting part of the cycle, any reverse current in T_1 is fed through T_e and the meter A_2. This current is made up of reverse-emission and ohmic-leakage currents through T_1. If a curve of reverse current against output current from T_1 is plotted, it will take the form shown in Fig. 36.

At low values of output current the reverse current consists entirely of the ohmic leakage current and tends to decrease in value with increase in output current.

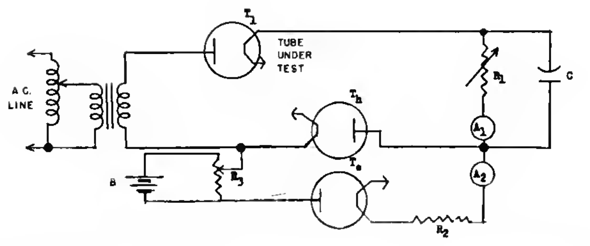

Fig. 35—Circuit arrangement for measuring reverse emission in diodes. The circuit may be duplicated for measurement of full-wave rectifier circuits.

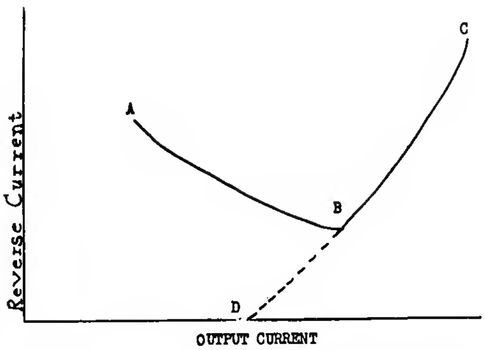

Fig. 36—Typical curve of reverse current vs output current in a rectifier.

The decrease occurs because the peak-inverse voltage across the tube under test is decreasing. As the temperature of the anode rises, the thermionic emission from it may increase rapidly, as shown in Fig. 36 by *CD*.

6.4 Primary Screen-Grid Emission

Measurement of the thermionic screen emission as a function of the power dissipation may be accomplished by use of the circuit shown in Fig. 37. By means of suitable rectifiers and a 60-cycle source, the screen is heated during the positive half-cycles and the primary emission is measured during the negative half-cycles. Appropriate direct anode voltage and control-grid bias are applied to the tube, and the screen dissipation is set to the desired level by adjusting the alternating input voltage of the transformer and reading current on meter A_1 and voltage on voltmeter V. T_h is a diode of very low impedance compared to the impedance of the screen circuit. The current read on meter A_2 is the primary screen emission (usually in the order of microamperes) under the conditions of test.

A wattmeter instead of voltmeter V and ammeter A_1 will serve to establish screen dissipation in this test without tedious calculation.

In some cases a simplified version of this circuit will give useful information, without the direct voltage supply and with the anode and the control grid tied to cathode.

6.5 Primary Anode Emission

Measurement of the thermionic emission from the anode as a function of dissipation may be accomplished

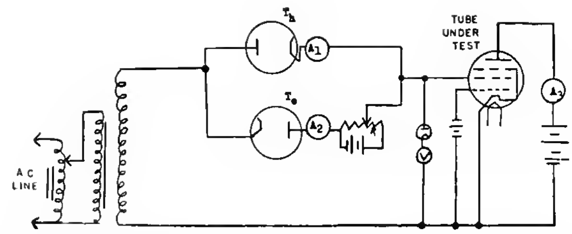

Fig. 37—Circuit arrangement for measuring primary emission from screen grid of a pentode.

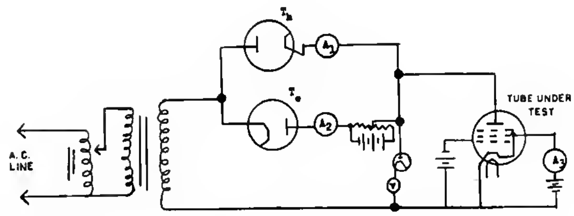

Fig. 38—Circuit arrangement for measuring primary emission from anode of a pentode.

by the circuit of Fig. 38, which is similar in principle to the circuit discussed in 6.4 for the screen emission measurement. The desired anode dissipation is obtained by applying alternating voltage to the anode and reading current on ammeter A_1 and voltage on voltmeter V. Appropriate values of screen and control-grid bias are applied. The primary anode emission is read on meter A_2

7. Vacuum-Tube Admittances

From the point of view of the circuit-design engineer, it is desirable to have a set of parameters that are characteristic of the tube alone, and that will enable a direct prediction of its behavior to be made over the useful frequency range when known or specified admittances are attached to its physically available terminals.

For zero frequency or very low frequencies up to the order of 10^3 cps, this specification and prediction of behavior may be accomplished through the measurement and use of the admittances comprising the familiar equivalent circuit that may be drawn for a triode as shown in Fig. 39.

For frequencies of the order of 10^3 to 10^6 cps, the effects of the input, output, and feedback capacitances are handled by adding them appropriately, as shown in Fig. 40.

In Section 7.1 methods are described for measuring the various capacitances, such as C_{gk}, C_{gp}, and C_{pk}. If these measurements are made directly at the base of the tube, it must be realized that only the internal electrode and lead capacitances have been determined. To make possible the prediction of the behavior when known terminations are attached to the available terminals, the capacitances of the socket and shielding must be added to these capacitances. This may be accomplished most directly by measuring the desired capacitances in the actual socket and shield configuration to be used. Some standard conditions for such measure-

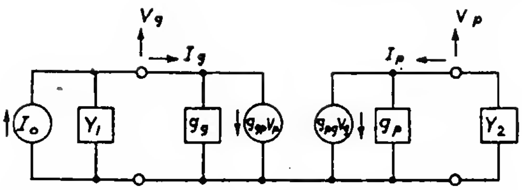

Fig. 39—Low-frequency equivalent circuit of a triode.

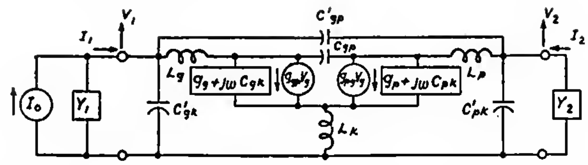

Fig. 41—Equivalent circuit of a triode for frequencies up to 100 Mc.

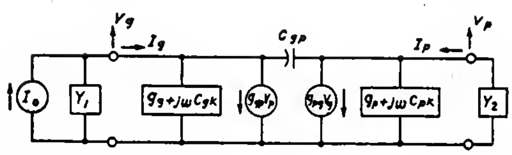

Fig. 40—Equivalent circuit of a triode for frequencies up to 1 Mc.

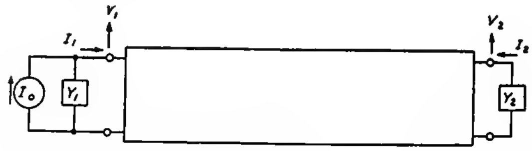

Fig. 42—Four-pole representation of an electron tube.

ments are given in the Electronic Industries Association standards. If an unconventional use of the tube is contemplated, the capacitances should be measured at the available terminals of whatever shield and connection arrangement is to be used.

In Section 7.2 methods are described for measuring the real parts of the various admittances making up the equivalent circuit, such as g_p, g_{pg}, g_{gp} and g_g. At frequencies up to about 10^6 cps, it is relatively unimportant whether these conductances are measured at the tube base or in the socket, since both the internal and external leads are sufficiently short so that only their capacitive effects need be considered.

At frequencies greater than about 10^7 cps, marked increases in the input, output, forward, and feedback admittances occur other than those caused by the grid-plate capacitance. The causes for these modifications can generally be traced to either or both of the following effects: 1) the effect of the passive coupling circuit (both internal and external) connecting the electron stream to the externally available terminals; 2) the effects of the finite transit time of the electrons through the inter-electrode spaces. When tubes with wire leads and closely spaced electrodes are used in conventional sockets at frequencies up to about 10^8 cps, these increases are caused primarily by the coupling networks connecting the electron stream to the external terminations. These effects are taken into account in Fig. 41 by indicating the lead self-inductance and mutual inductance and splitting the low-frequency capacitances into electrode parts and circuit parts. Not only has the circuit now become complicated for analysis, but a more serious defect has appeared: the new circuit elements that have been introduced cannot be measured directly from the external terminals alone.

At frequencies higher than about 10^8 cps, additional modifications in the basic electronic admittances caused by transit time make it still more difficult to assign values to the various circuit elements either by measurement or by calculation.

If we ignore the internal electronic and circuital complexity existing between the available input and output terminals and focus attention only upon these terminals, we have simply the four-terminal box of Fig. 42. As is well known, four parameters are necessary to specify completely the behavior of such an active linear transducer. From among the many sets of such parameters that are available, consider the short-circuit input admittance y_{11}, the short-circuit output admittance y_{22}, the short-circuit forward admittance y_{21}, and the short-circuit feedback admittance y_{12}.

The behavior at all frequencies is then indicated by the nodal equations (1) and (2) subject to the terminal conditions (3) and (4). (See equations below.)

$$I_1 = y_{11}V_1 + y_{12}V_2 \tag{1}$$

$$I_2 = y_{21}V_1 + y_{22}V_2 \tag{2}$$

$$I_0 = I_1 + Y_1V_1 \tag{3}$$

$$I_2 = -\ Y_2V_2 \tag{4}$$

$$Y_{11} = \frac{I_1}{V_1} = y_{11} - \frac{y_{12}y_{21}}{y_{22} + Y_2} \tag{5}$$

$$\frac{V_2}{V_1} = \frac{-y_{12}}{y_{22} + Y_2}\,. \tag{6}$$

Solution of these equations leads to expressions for the various design quantities, such as input admittance and voltage amplification, which will be valid for all frequencies, only the values of the admittances changing with frequency. These admittances may be arranged in a simple circuit, as shown in Fig. 43. Comparison with Fig. 39 makes it apparent at once that the circuits are the same in configuration so that the short-circuit admittances do reduce to the familiar electron-tube coefficient at low frequencies.[3]

As the frequency is increased, these admittances will vary and reflect in their changing values the effects of the coupling circuits and electron-transit times.

The important fact to be emphasized is that these ad-

[3] Associated with the admittance type of analysis must be the concept of impressed currents. There are two graphical symbolisms for indicating the introduction of impressed currents into a circuit. These are shown in Figs. 44 and 45. That shown in Fig. 44 is used in all the preceding figures.

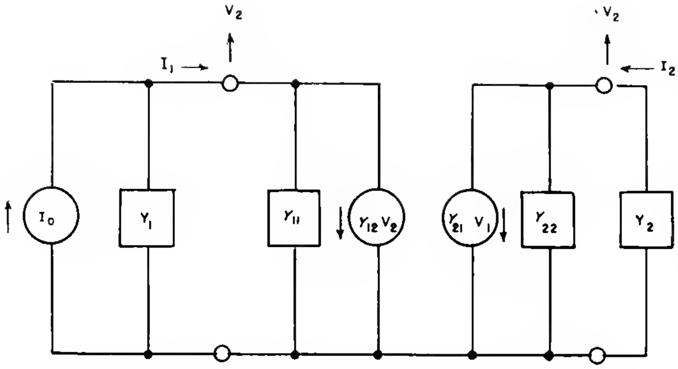

Fig. 43—Equivalent circuit of an electron-tube four-pole.

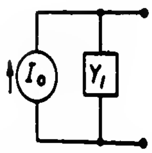

Fig. 44—Symbolism (a) for indicating the introduction of an impressed current.

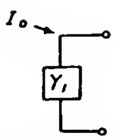

Fig. 45—Symbolism (b) for indicating the introduction of an impressed current.

mittances can be determined from measurements made only upon the external terminals, whereas the elements of Fig. 41 cannot be so determined.

In Section 7.3, methods are described in principle for measuring the short-circuit admittances over a wide frequency range. Apparatus employing lumped-circuit and coaxial-line-circuit techniques is described in some detail for frequencies of the order of 10^7 to 10^9 cps. By using the same principles, but with different physical configurations of the circuits such as waveguides with standing-wave devices, the frequency range may be extended indefinitely.

7.1 Direct Interelectrode Capacitances

It is recommended that direct capacitances be measured, rather than total capacitances, each of which is the sum of two or more direct capacitances. In measurements of direct capacitances between two elements of multi-element tubes, it is customary to connect all other elements to ground unless another connection is desired which more nearly simulates circuit operating conditions. Published capacitance values usually specify connections for the other elements. In general, it is customary to connect the heaters and screen grids to the cathode and to connect other sections or units to ground. Information as to the element connections for specific published capacitance values should be obtained from the published data sheets. The three direct capacitances of a triode are grid-plate capacitance C_{gp}, grid-cathode capacitance C_{gk}, and plate-cathode capacitance C_{pk}. When a tube is active, its direct interelectrode capacitances differ from the values obtained with a cold tube. The difference may be sufficient to be of importance in certain applications.

Interelectrode capacitances, unless otherwise specified, are measured with the cathode cold and with no direct voltages present.[4]

There follow several methods of measurement for interelectrode capacitances. The radio-frequency bridge method and the transmission method are applicable throughout the usual ranges of tube capacitances of the order of 0.0001 to 100 pf. Substitution Method A is useful for capacitances above 1.0 pf. Substitution Method B is more suitable for smaller capacitances. The capacitance values as read will depend on shielding geometry.

Audio-frequency bridge measurements are not currently used for the measurement of small values of tube capacitances.

7.1.1 Radio-Frequency Bridge Method: A bridge circuit for the measurement of direct interelectrode capacitances of a tube is shown in Fig. 46. A stable oscillator, such as a crystal-controlled oscillator, supplies RF power through a closely coupled balanced transformer (T). Balance is indicated by a null-indicating vacuum-tube voltmeter which is made up of a tuned amplifier, diode rectifier, and direct-current meter indicator. For convenience, the capacitors are ganged differentially so that increase of one capacitance is accompanied by an equal decrease of the other. Balance may then be effected by varying the two capacitance branches until they are equal (when $C_x = C_1 - C_2$). Then, at balance $C_x = |2\Delta C_1| = |2\Delta C_2|$. When an adapter is used, the tube capacitance is the difference in capacitance readings with the tube in and out of the adapter. An advantage of the bridge over the transmission or the substitution method is that the conductive components of the tube admittances due to insulation losses, getter deposits, or other leakages, can be measured and balanced out independently of the capacitance reading. The effect of capacitance to ground is negligible as point B is at center location in the bridge, where capacitance does not influence balance; and the capacitance from C to ground is across a closely coupled low-impedance winding that does not affect the capacitance balance or the voltage supplied to the bridge.

7.1.2 Transmission Method: A circuit for measuring the direct interelectrode capacitances of a tube is shown schematically in Fig. 47. The RF oscillator voltage is attenuated according to the range desired. The current in the unknown tube capacitance is amplified and measured by a tube voltmeter. The amplifier input is attenuated concomitantly in conjunction with the os-

[4] See Standard RS-191 of the Electronic Industries Association (formerly the Radio-Electronic-Television Manufacturers Association) for details of approved conditions of measurement.

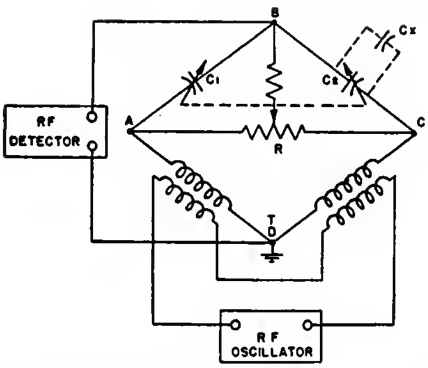

Fig. 46—Circuit arrangement for RF bridge for measuring interelectrode capacitances.

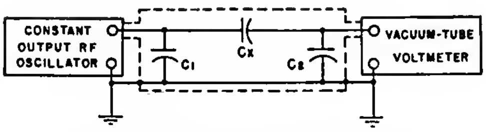

Fig. 47—Circuit arrangement for measuring interelectrode capacitance by transmission method.

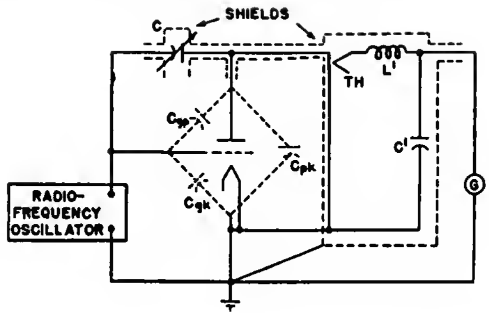

Fig. 48—Circuit arrangement for measuring interelectrode capacitance by Substitution Method A.

cillator output so that various ranges may be obtained. The oscillator-output and amplifier-input attenuators may be operated from a common control and calibrated in convenient decade steps. It is to be noted that large capacitors are required across the input and output so that the effect of the tube capacitances shunted across the input and output is negligible. The device is calibrated by using a known standard capacitor or a resistor of negligible shunt capacitance that may be calibrated in place. The parts must be shielded from one another to eliminate stray capacitance because there is no way of balancing them out with this method.

7.1.3 Substitution Methods:

7.1.3.1 Method A, substitution of calibrated capacitor for tube capacitance: Fig. 48 illustrates a substitution method (with set-up connected for measurement of grid-plate capacitance C_{gp}) using a radio-frequency oscillator as the source and a thermoelement *TH* and galvanometer *G* as the indicator. Alternatively, a vacuum-tube voltmeter may be used as an indicating device to obtain higher sensitivity. The shielded capacitor *C* is calibrated to read capacitance above an arbitrary reference point and should have a range as great as the largest capacitance to be measured. With *C* set at this reference point and the vacuum tube in the circuit, the galvanometer reading is taken. The vacuum tube is removed and *C* is adjusted until the galvanometer reading is the same as before. The added capacitance of *C* is then equal to the grid-plate capacitance C_{gp}. The RF oscillator should maintain a constant product of voltage and frequency. To verify that the oscillator is maintaining a constant product of voltage and frequency, a thermoelement with galvanometer and filter may be connected in series with a small capacitance across the oscillator terminals. The connection from the plate of the vacuum tube through the thermoelement *TH* to the filament should be short if it is not shielded. The capacitor *C′* and coil *L′* constitute a filter system to keep RF current from flowing through the lead to the galvanometer *G*.

7.1.3.2 Method B, substitution of calibrated capacitor for low-tube capacitances: For small capacitances such as the grid-plate capacitance of screen-grid tubes, the substitution method of Fig. 49 can be employed with a calibrated capacitor *C* of suitable range, a RF source of suitable range, and a detector of somewhat greater sensitivity than a thermoelement.

Because of the great sensitivity required to measure the very small values of capacitance, it is desirable that all disturbing influences be minimized. This is accomplished by keeping the capacitances across the oscillator and across the detector constant by means of a balancing tube.

The low-capacitance switch *S* is first thrown to the tube T_1 under test and the reading of the microammeter noted. The switch is then thrown to T_2, the balance tube, which should be of the same type as T_1, and the capacitor *C* is adjusted to give the same reading of the microammeter as before. The feedback capacitance C_{gp} is then equal to the added capacitance of *C*.

7.1.3.2.1 Precautions: In Method B the voltage and frequency of the oscillator should remain constant, as indicated on the microammeter when a capacitive load is added. If this is not possible, the product of its output voltage and frequency must be constant.

Fig. 49 shows a balance tube T_2. If such a balance tube is used, its input resistance should be greater than 100 megohms; otherwise, serious errors will result when low values of grid-plate capacitance are measured. If a tube with this high input resistance is not available, a calibrated low-loss capacitor having a capacitance within 20 per cent of the grid-cathode capacitance of T_1 may be used instead.

7.1.3.3 Method C, comparison of input voltages for standard and unknown capacitances for constant output: An alternative substitution method of high sensitivity is illustrated in Fig. 50. This method employs a calibrated variable-voltage source and a fixed standard capacitor instead of the fixed-voltage source and cali-

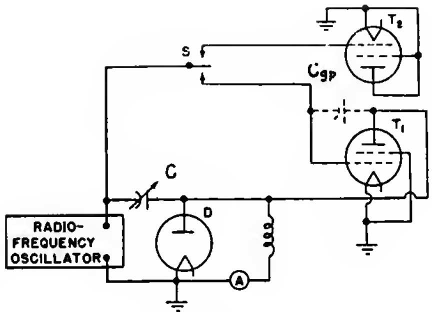

Fig. 49—Circuit arrangement for measuring interelectrode capacitance by Substitution Method B.

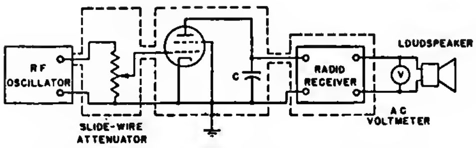

Fig. 50—Circuit arrangement for measuring interelectrode capacitance by Substitution Method C.

brated variable capacitor of Substitution Methods A and B. A standard signal generator is convenient for this purpose. The current through the grid-plate capacitance of the tube produces a voltage across the receiver input, which is shunted by a large capacitance C. In the arrangement shown in Fig. 50 the RF oscillator is modulated, and an ac voltmeter connected across the terminals of the loudspeaker is used as an indicator.

A standard fixed capacitor C_s of the order of one-half pf, suitably shielded, is first used in place of the tube shown in Fig. 50, and the attenuator is set so that the impressed RF voltage is E_s with a standard audio-frequency output. The tube to be measured is next substituted for the standard capacitor and the attenuator readjusted until an impressed RF voltage E_x produces the same deflection at V. The unknown grid-plate capacitance is

$$C_{gp} = \frac{C_s \quad E_s}{E_x} \, .$$

The standard capacitor can be enclosed within a vacuum tube of standard dimensions. Two circular disks, 2 cm in diameter and spaced 8 mm apart, will provide a standard capacitor of the proper order of magnitude. It may be calibrated by measurements on a differential capacitance bridge.

7.1.3.3.1 Precautions: The leads from the attenuator to the tube shield and from the latter to the voltage-measuring unit should be completely shielded, and the tube itself should be enclosed in a rather closely fitting cylindrical shield. For a double-ended tube, the test set is provided with a small contactor to the grid (or plate) cap, the lead to this contactor entering at the top of the shield. The lead from the plate or grid leaves at a point near the bottom of the shield. For a single-ended tube, shielding adequate for the elimination of capacitance components between the active tube and socket terminals should be provided. The adequacy of the shielding may be checked by readings made with the tube removed from the socket, unless the tube base itself provides a shield.

The shunt capacitance C should be large compared to the plate-cathode capacitance of the tube under test; ordinarily a value of 500 pf is satisfactory.

7.2 Vacuum-Tube Coefficients

This section is devoted to methods for measuring the low-frequency coefficients of vacuum tubes. The more commonly used coefficients are plate resistance, grid-plate transconductance, amplification factor, and conversion transconductance. The methods outlined apply not only to these coefficients, but also to the less commonly used coefficients referred to any one or to any pair of electrodes, such as plate-grid transconductance, grid resistance, conductance for rectification, grid-screen μ factor, etc.

In general, most tube coefficients may be evaluated from the characteristic graph (see Section 4), or may be measured directly by a balance or null method of measurement employing an audio-frequency generator as the source of power and a null indicator (usually a telephone receiver, which is preceded by an amplifier for more precise results). The results obtained by balance methods are usually more reliable than the results obtained from the characteristic graphs, particularly in the measurment of coefficients of coated-cathode tubes. In the itemized portion of this section various coefficients are discussed individually. The basic circuits for measuring the coefficients by balance methods are shown schematically in Figs. 51 to 59. The circuit for measuring conversion transconductance is shown in Fig. 61.

Two types of balance methods are described in this section. Figs. 51–54 (and in more detail, Fig. 55) illustrate the resistance-ratio type of circuit in which an alternating voltage is introduced at only one point in the circuit, a null is obtained by the adjustment of impedances, and the value of the coefficient is expressed by a resistance ratio. Figs. 56–58 (and in more detail, Fig. 59) illustrate the voltage-ratio type of circuit in which two or more properly phased alternating voltages are introduced into the circuit, a null is obtained by adjusting their relative magnitudes, and the value of the coefficient is expressed by a voltage ratio.

The voltage-ratio method uses the same component parts for measuring all of the coefficients. Details of the method are illustrated in Fig. 59 which is an expanded version of Fig. 58 as applied to the determination of the amplification factor of a pentode. The method utilizes a system of transformers, capacitors, and attenuators to supply three independent, properly phased and adjusted voltages from a common source. The value of the coefficient at balance is a voltage ratio as indicated by

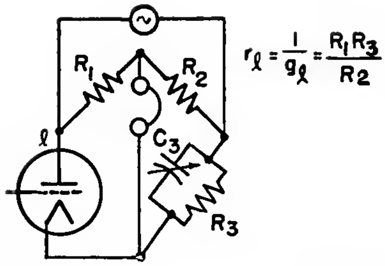

Fig. 51—Basic resistance-balance circuit for measuring electrode resistance.

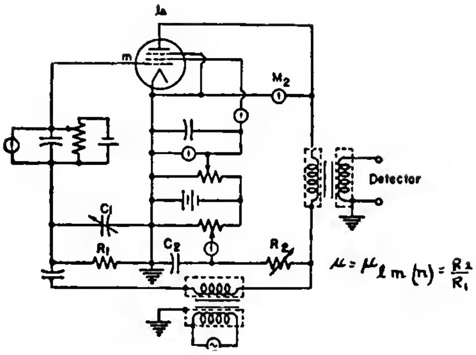

Fig. 55—Detailed resistance-balance circuit for measuring μ factor.

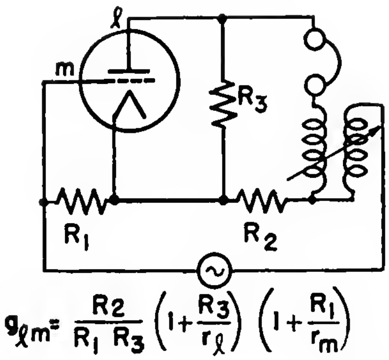

Fig. 52—Basic resistance-balance circuit for measuring transconductance.

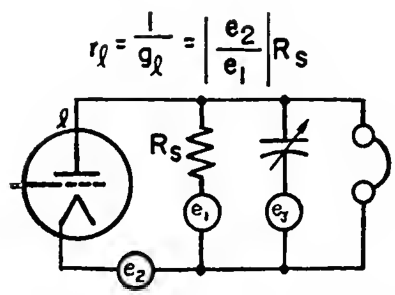

Fig. 56—Basic voltage-ratio circuit for measuring electrode resistance.

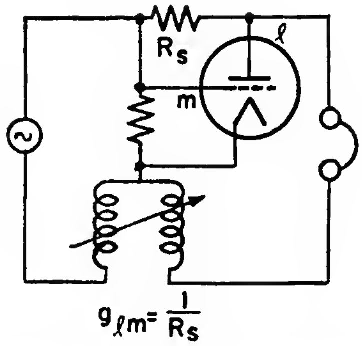

Fig. 53—Basic resistance-balance circuit for measuring transconductance of tubes having low plate resistance.

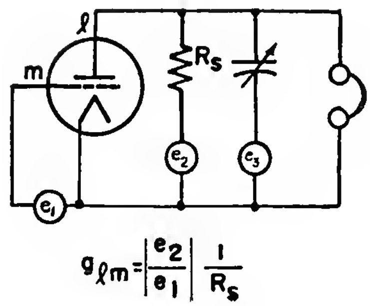

Fig. 57—Basic voltage-ratio circuit for measuring transconductance.

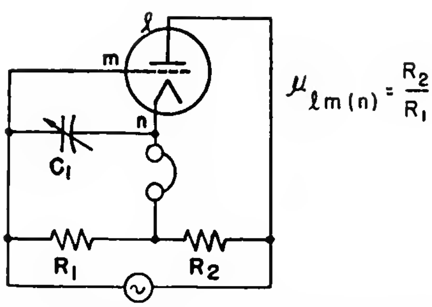

Fig. 54—Basic resistance-balance circuit for measuring μ factor.

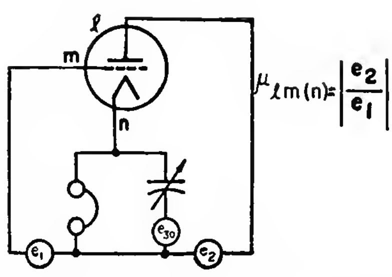

Fig. 58—Basic voltage-ratio circuit for measuring μ factor.

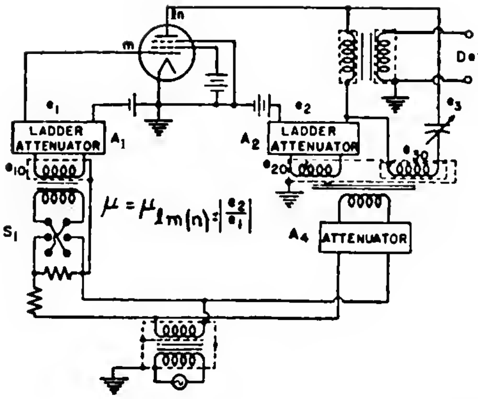

Fig. 59—Detailed voltage-ratio circuit for measuring μ factor.

the attenuator settings (Fig. 59), where the attenuator A_4 indicates the significant figures, and where the attenuator A_1 or the attenuator A_2 indicates the decimal point. The voltages e_{10}, e_{20}, and e_{30} must be in phase. This is accomplished by using similar transformers. Since the output resistances of the attenuators at e_1 and e_2 may be made small by proper design (less than 25 ohms), the voltage drop due to electrode currents through these attenuators usually may be neglected, and the electrode voltages are unaffected by the balancing procedure. Since the voltage sources are insulated from one another, the batteries or power supplies may be grounded at the cathode, and the phase-balancing voltage e_3 may always be connected across the null detector. The reversing switch S_1 permits measuring negative coefficients.

7.2.1 Electrode Resistance and Electrode Conductance:

7.2.1.1: The electrode resistance and electrode conductance may be obtained from the graph of the current to the electrode as ordinate, plotted against the voltage between that electrode and the cathode, all other electrode voltages being maintained constant. The tangent to this characteristic at any point is the electrode conductance at the electrode voltages corresponding to this point. The reciprocal of the tangent is the electrode resistance.

7.2.1.2: Electrode resistance and electrode conductance may also be determined by the bridge method given in Fig. 51. At balance the resistance and conductance are given by the relation

$$r_l = \frac{1}{g_l} = \frac{R_1 R_3}{R_2}\,.$$

In bridge measurements the quadrature current is usually best balanced out by a small capacitor in an arm of the bridge adjacent to the arm containing the unknown. Care should be taken in choosing the bridge-arm resistances to insure that the current flowing to the electrode does not vary appreciably while the bridge is being balanced to a null.

7.2.1.3: The voltage-ratio circuit for determining the electrode resistance and electrode conductance is given in Fig. 56. At balance the values of the coefficients are as follows:

$$r_l = \frac{1}{g_l} = \left|\frac{e_2}{e_1}\right| R_s.$$

R_s is a fixed resistance of perhaps 100,000 ohms, and the voltage ratio is indicated by the attenuator settings.

Grid conductance and plate conductance, as well as grid resistance and plate resistance, are particular electrode coefficients and are measured by any of the methods outlined above.

7.2.1.4: Another method is available which lends itself to rapid determination of the plate resistance r_p of tubes such as the screen-grid type, where the value of plate resistance is one-half megohm or higher. The circuit arrangement is shown in Fig. 60.

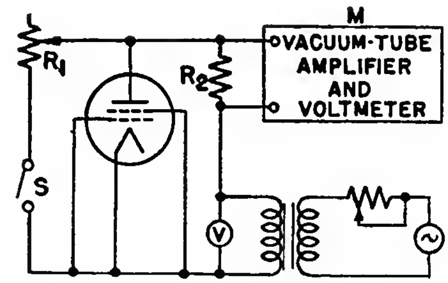

Fig. 60—Basic circuit for measuring the plate resistance of screen-grid tubes by the substitution method.

The tube is operated at normal voltages. Switch S is open. Alternating voltage is applied as indicated and adjusted until a convenient deflection is obtained on the vacuum-tube voltmeter M. The tube is then removed from the circuit, and a resistor R_1 substituted for the internal resistance of the tube by closing switch S. R_1 is then adjusted to give the same deflection on M with the alternating voltage held constant at its original value. The value of R_1 is then the value of the plate resistance of the tube. R_2 should be made negligibly small in comparison with the plate resistance of the tube.

The arrangement can be made into a direct-reading device in the following three ways:

a) By maintaining the alternating voltage constant and calibrating M in terms of the plate resistance of the tube. R_1 can be used in the manner described above for making calibration.

b) By maintaining the deflection of M constant and calibrating the alternating-current voltmeter in terms of the tube plate resistance, R_1 being used for calibration purposes. This method gives a straight-line calibration between the alternating voltage and the tube plate resistance and is more conveniently used than a).

c) By maintaining the alternating voltage constant and varying R_2 to give constant deflection of M. Then r_p will be proportional to R_2.

7.2.2 Transconductance:

7.2.2.1: Transconductance between any two electrodes may be determined graphically from the tangent to the graph of the current to the second electrode as ordinate, plotted against the voltage on the first electrode as abscissa, all other electrode voltages being maintained constant.

7.2.2.2: Transconductance may also be measured directly by the balance method of Fig. 52.

$$g_{lm} = \frac{R_2}{R_1 R_3}\left(1 + \frac{R_3}{r_l}\right)\left(1 + \frac{R_1}{r_m}\right).$$

If R_3 and R_1 are negligibly small compared with the resistances of the electrodes l and m, the balance equation reduces to

$$g_{lm} = \frac{R_2}{R_1 R_3}\,.$$

7.2.2.3: A second balance method, which is useful for measuring power tubes, or in general when rl is small, is shown in Fig. 53. The null is obtained by adjusting R_s, and at balance the transconductance is

$$g_{lm} = \frac{1}{R_s}\,.$$

The value of the resistor in the grid circuit does not affect the measurement.

7.2.2.4: The voltage-ratio method, which may be used for measuring transconductance of any value, is shown in Fig. 57. Here the transconductance is

$$g_{lm} = \left|\frac{e_2}{e_1}\right|\frac{1}{R_s}\,,$$

where R_l is usually 100,000 ohms, and the voltage ratio is indicated by the attenuator settings.

The grid-plate transconductance, also known as the mutual conductance ($g_{pg} \equiv g_m$), may be measured by any of the methods outlined above.

7.2.3 μ Factor:

7.2.3.1: The μ factor may be measured dynamically or statically. The method of making static measurements is indicated by the defining equation

$$\mu_{lm(n)} = \left|\frac{\Delta e_l}{\Delta e_m}\right| i_n = \text{const.}$$

7.2.3.2: Dynamically, the μ factor may be measured by the balance method of Fig. 54. The μ factor $\mu_{lm(n)}$ (the relative control of voltages at electrodes l and m on the current to electrode n) is determined by inserting the null indicator into the lead to electrode n. At balance,

$$\mu_{lm(n)} = \frac{R_2}{R_1}\,.$$

7.2.3.3: An alternative dynamic-balance method for measuring the μ factor is illustrated in Fig. 58. At balance,

$$\mu_{lm(n)} = \left|\frac{e_2}{e_1}\right|.$$

7.2.3.4: The amplification factor of a tube is that μ factor for which the plate of the tube is both the constant-current electrode and one of the electrodes involved in relative control effect of voltage. Therefore, in determining the amplification factor the circuit of either Fig. 54 or 58 may be used, but the null detector must ordinarily be placed in the plate circuit (as shown in Figs. 55 and 59), since the defining equation states that the alternating component of the plate current is zero. If the tube is a triode and if there is no grid current flowing, the null detector may be placed in the cathode circuit.

7.2.4 General Precautions for Balance Methods: The magnitude of the impressed alternating voltage should always be small enough so that the results of the measurement are unaffected by a reduction of the impressed voltage.

All electrodes not directly involved in the measurement must be maintained constant at specified voltages.

To indicate the extent of circuit modification sometimes required to eliminate sources of error, the basic circuit for measuring μ factor by the resistance-ratio method (Fig. 54) is shown expanded in Fig. 55. Since this figure shows a method for measuring the amplification factor of a pentode (see paragraph 7.2.3.2), the detector is in the plate circuit. The impedance of C_2 must be negligible in comparison with resistance R_2 at the test frequency.

Balance methods employing an alternating-voltage generator require that consideration be given to the effects of stray capacitances and couplings, which may render balance difficult or impossible. The grounding and shielding of apparatus should be given special attention.

Batteries and, in particular, dc supplies operated from the power line may introduce excessive capacitance across the network elements unless properly located with respect to ground. In the example of Fig. 55, the plate and screen supplies are at ground potential and the capacitance of the grid supply may be neglected because of the low resistance of R_1. Capacitors of sufficiently low impedance to by-pass the audio-frequency currents must usually be shunted across the power supplies. Voltage-regulated power supplies are excellent low-impedance devices for this purpose.

Audio-frequency generators and amplifiers that are operated from the power line will have appreciable capacitance to ground, the effect of which must usually be minimized by the use of shielded transformers, as in Fig. 55, or of guard circuits.

When direct current must flow through the generator

or through the detector, low-resistance transformers or chokes must be used to minimize the direct voltage drop. Spurious coupling between the chokes or transformers must then be prevented by proper shielding, orientation, and spacing of the component parts.

In some circuits the direct voltage at an electrode may vary as balance adjustments are made. When this cannot be avoided by means of a blocking capacitor or by circuit rearrangement, the direct voltage is best measured at the electrode (see M_2 of Fig. 55) with a meter of sufficiently high resistance to preclude errors due to its shunting effect.

Circuit resistors must be able to carry the electrode currents with negligible effect on their resistance value.

In the measurement of the coefficients of high-transconductance tubes, oscillation may occur. Oscillation of the tube under test may be evident in the lack of a sharp null point or may not be evident except in erroneous readings. Checks by other methods should therefore be made to establish freedom from oscillation in setting up a new bridge or in measuring high-transconductance tubes. Oscillation can be checked by the presence of grid current or by the change of plate current when the grid or plate terminal is grounded for the spurious frequency. Oscillation may be prevented by exercising care in choosing the length and the location of the leads to the tube socket, by the judicious use of by-pass capacitors, or by the insertion of a resistor in series with the appropriate electrode.

Circuit and interelectrode capacitances introduce quadrature currents that may have to be balanced out to attain an exact null. Various methods of quadrature balance are available. The mutual-inductance method illustrated in Fig. 53 couples, in proper phase, voltage from the generator to the detector (or, on occasion, to the grid circuit, if there is no grid current). Because of the resistance, self-inductance, and mutual inductance introduced in series with the detector, the balance conditions are modified, and a slight error may exist. In the capacitance method a small variable capacitor is used, as for example at C_1 in Figs. 54 and 55. Both of these methods require readjustment of the circuit for every change in the main balance network. This difficulty may be avoided by using an auxiliary capacitance bridge to balance out the quadrature component before the tube filament is turned on. In the voltage-ratio method the out-of-phase voltage at the detector is balanced out by an equal voltage of opposite phase connected in parallel with the detector terminals.

7.2.6 Conversion Transconductance (Conversion Conductance):

7.2.6.1 Low-frequency method: The definition of conversion transconductance shows that if $f_1 = f_2$, the difference frequency will be zero, corresponding to direct current. If the two signals are in phase, the plate current will be higher than with the two signals 180° out of phase. The difference in direct plate current, caused by a reversal of the phase relation of the two signals, divided by twice the peak value of the alternating voltage on the signal grid, is the conversion transconductance. It is assumed that the voltage on the oscillator grid is adjusted to such a value that the current drawn by the grid corresponds to its rated value.

In practice, measurements may be made on multigrid converter and mixer tubes in a circuit such as that shown in Fig. 61. With grid resistor R_1 and grid current as rated, a signal voltage E_g is applied to the signal grid 180° out of phase with the voltage applied to the oscillator grid. The bucking circuit is adjusted to give zero reading of the plate-current meter, after which the phase of the signal voltage applied to one of the grids is reversed and the reading of the plate-current meter ΔI_b noted. Conversion transconductance is then

$$g_c = \frac{\Delta I_b}{2\sqrt{2}\, E_g}\,.$$

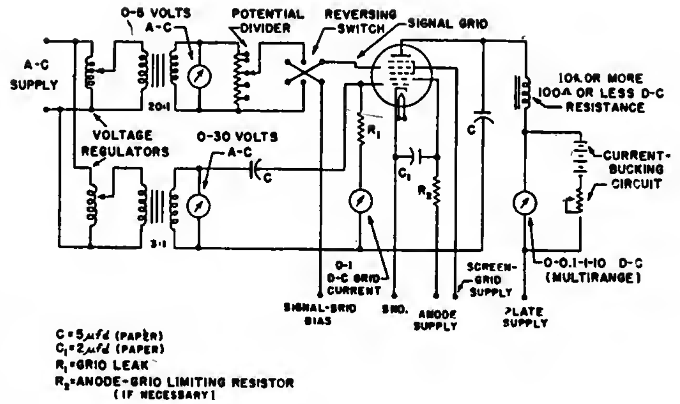

Fig. 61—Circuit arrangement for measuring conversion transconductance.

For precision measurements, the signal voltage applied to the signal grid should be held to as low a value as possible, and in no case should cause current to flow in the signal-grid circuit. For most converter tubes, it is convenient to make E_g equal to 0.354 volt rms. The conversion transconductance in micromhos is then equal to ΔI_b in microamperes.

In the circuit of Fig. 61 it should be noted that the transformers and regulators must have sufficiently low leakage reactance and resistance to maintain exact phase relation between the signal and grid-driving voltages. The regulators may be variable autotransformers or the equivalent. The resistance of the bucking circuit must be high compared to the resistance of the plate-current meter. (A ratio of 10:1 will produce an error of about 10 per cent, and 100:1 will result in an error of about 1 per cent.)

The following procedure must be followed in making conversion-transconductance measurements when a screen-voltage dropping resistor is included in the test circuit to simulate the normal circuit with the resistor

by-passed: 1) With the screen resistor in the circuit, adequately by-passed for 60-cycle current, and with zero signal on the signal grid and normal oscillator-grid current, read the screen current. 2) Remove the screen resistor and apply a voltage to the screen equal to the supply voltage minus the voltage drop computed from the screen current determined above. Then read conversion transconductance in the usual manner.

When a tube is to be operated with grid-resistor bias, conversion transconductance must be measured with a separate bias adjusted to give the same plate current as would be obtained with the grid-resistor bias.

When a tube is to be operated with a cathode-bias resistor included in the test circuit to simulate the normal circuit with the resistor by-passed, the following procedure should be used: 1) With the cathode resistor in the circuit adequately by-passed for 60-cycle current, and with the correct oscillator-grid current and specified grid resistor, the individual currents are read. 2) The cathode resistor is removed or short-circuited and bias is applied to the control grid. The control-grid and screen voltages are adjusted to give the same electrode currents as before. The conversion transconductance is then read in the usual manner for each value of bias resistance. In the measurement of the conversion transconductance of pentodes and triodes with signal-grid injection, both the ac signals shown in Fig. 61 are applied to the signal-grid.

7.2.6.1.1 Precautions: The 60-cycle voltage must be reasonably sinusoidal. Voltage regulators of the saturation type should be avoided because of waveform distortion.

The regulation of the screen supply should be such as to have a negligible effect upon the plate current, regardless of the phase of the signal voltage. For example, the use of a screen source having a dc resistance of 10,000 ohms in measurements on ordinary converter tubes may cause plus or minus 0.6 per cent voltage variation with a resultant error in conversion transconductance reading of over 15 per cent.

7.2.6.2 High-frequency method: A circuit for this method is shown in Fig. 62. Values of circuit constants are included in Fig. 62 for illustrative purposes only. An input voltage of 50 mv with a frequency of 10 kc is used, and an appropriate 11-kc voltage from a separate oscillator is applied to the circuit. The product of the resistance of the oscillator grid multiplied by the oscillator coupling capacitance C_{g3} should be 2×10^{-4} ohm-farads. The anode circuit is tuned to 1 kc and has a resistance of 50 ohms. The voltage of the tuned circuit is measured with a tube voltmeter, and the conversion transconductance is calculated from this voltage, the impedance of the tuned circuit, and the 10-kc input voltage.

7.2.7 Oscillation Test for Converter Tubes: The oscillator performance of a converter tube depends upon several tube and circuit parameters that are not simply defined. The test-oscillator circuit shown in Fig. 63,

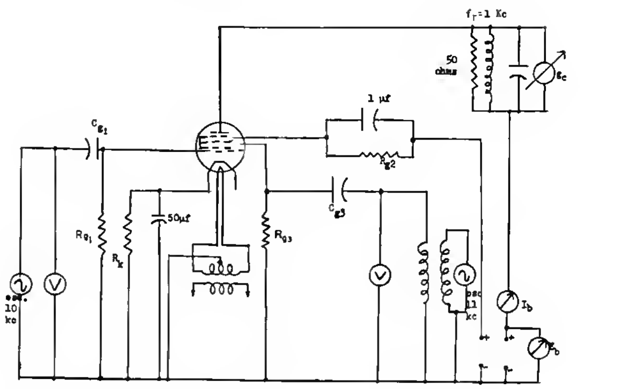

Fig. 62—Circuit arrangement for measurement of conversion transconductance g_c at 10 kc. Values of circuit constants are for illustrative purposes only.

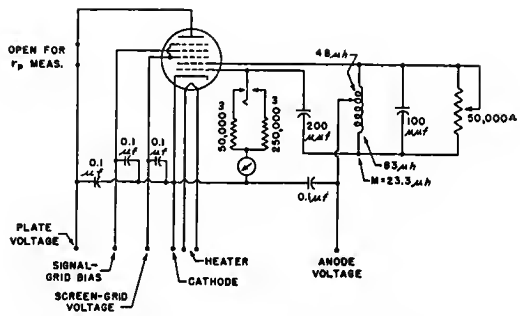

Fig. 63—Converter-tube test oscillator.

while not giving an exact criterion of oscillator performance, does provide means for determining the usual parameters under conditions that simulate average service for composite converters. It employs a Hartley circuit having a fixed feedback ratio and having the tank-circuit impedance variable for adjustment of the magnitude of oscillation.

Bias is obtained by means of a capacitor and a wire-wound gridleak. The wire-wound grid resistor is used because it has considerably higher impedance to radio frequencies than to direct current. The high RF impedance diminishes the shunting effects of the gridleak on the tuned circuit. The rectified grid current is read on a microammeter in series with the grid resistor. Before the test oscillator is used, the relationship between the tuned impedance of the tank circuit and the setting of the variable resistor must be determined.

The ability of a tube to oscillate when the shunt impedance of a tank circuit is low is one criterion of the value of a tube as an oscillator. The oscillation test is made by applying the desired electrode potentials to the tube under test and reading the rectified grid current at some known setting of tank-circuit impedance. The minimum tuned impedance at which oscillation will start or cease, as indicated by rectified grid current, may be included as part of the test data.

7.2.7.1 Precautions: The constants for this circuit as given in Fig. 63 have been selected to simulate average circuit conditions in broadcast receivers for com-

posite converters having transconductance less than 1500 μmhos at zero bias. For tubes having higher transconductances or for circuits for use above broadcast frequencies, the circuit with the constants given may not be satisfactory, as spurious oscillations may make it impossible to obtain correlation between test readings and receiver performance.

This circuit in any form may be used to determine relative values. When comparison of values with readings from other sources is to be made, calibration is advisable.

7.2.8 Converter Plate Resistance: The plate resistance of a converter or mixer tube may be measured under conditions simulating operation. With suitable oscillator excitation and all direct potentials applied to the tube, the plate resistance is measured as described in Section 7.2.1.

When the plate resistance of a converter is measured by a LF bridge, use of a HF oscillator signal (2 to 7 Mc) may prove to be useful, because there is generally sufficient decoupling in the bridge itself to prevent the RF signal from upsetting the balance. When measuring a high impedance (*i.e.*, over 5 megohms) it is necessary to provide additional decoupling, but only capacitors of low value (not exceeding 50 pf) should be used, otherwise it may be impossible to balance the bridge.

7.3 Four-Pole Admittances

The definitions of the four-pole admittances indicate that the short-circuit driving-point and transfer admittances are to be measured with the tube operating. Since these admittances depend upon the shielding, socket, by-pass capacitors, and dc filters, as well as upon the internal circuit features and electron dynamics of the electron tube, it is important that the conditions of use be given as a part of the data until standard conditions for measurement are set up. When a new or nonstandard method of connection to the tube is contemplated, the admittances taken at the available terminals of the new arrangement may be markedly different from those obtained with the same tube in a different socket-filter combination.

Eq. (5) of Section 7 and the definition show that the short-circuit input admittance y_{11} may be obtained from measurement of the input admittance when the output termination Y_2 is effectively short-circuited at the frequency of measurement. When Y_2 is short-circuited, the second term of (5) is negligible. (Since a complete short circuit is not possible when the tube is operating, the effectiveness of the short circuit should be checked at the frequency of measurement by making certain that negligible variation in the input admittance is produced by an appreciable change in the output termination.) Similarly, the short-circuit output admittance may be obtained from a measurement of the output driving-point admittance when the input is effectively short-circuited at the frequency of measurement.

Bridge methods are preferable at all frequencies; however, two methods that have been used at high frequencies are given below.

7.3.1 Susceptance-Variation Method of Measurement: The following method of measurement is a form of the well-known reactance-variation method widely used for the measurement of two-terminal admittances. It is adapted to the determination of the transfer admittance, as well as to the driving-point admittances.

Fig. 64 is a semischematic diagram of the test equipment. In this figure T is the active or passive transducer to be measured and Y_1 and Y_2 are calibrated variable-admittance elements, which may be of various forms, such as coils and capacitors, or adjustable-length lines. Signal-frequency voltage-measuring devices V_1 and V_2 are placed across the input and output terminals of the transducer; these may be simply crystal or diode voltmeters or heterodyne receivers.

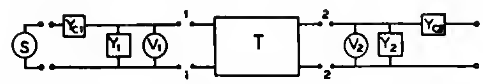

Fig. 64—Semischematic diagram of equipment for measuring admittances by the susceptance-variation method.

Variable admittances Y_{C1} and Y_{C2} are used for coupling the input or output circuits to the signal oscillators.

If all four admittances are to be measured at a single frequency, the following sequence of operations is recommended.

7.3.1.1 Measurement of y_{11} (short-circuit input admittance):

a) Short-circuit the output terminals 2-2. This may be done either by detuning Y_2 sufficiently or by placing a suitable by-pass capacitor directly across terminals 2-2.

b) Excite the input circuit by coupling the signal oscillator loosely through Y_{C1} to Y_1.

c) Adjust Y_1 for resonance as indicated by a maximum reading of V_1. In order to insure that the coupling to the oscillator is sufficiently small, reduce the coupling until further reduction does not change the setting of Y_1 for resonance. Record the calibrated values of G_1 and B_1 for this setting.

d) Vary Y_1 on either side of resonance until the voltage V_1 is reduced by a factor $1/\sqrt{2}$. Record the calibrated values of this total variation of Y_1 between half-power points as ΔG_1 and ΔB_1. In order to insure that the oscillator and detector are not loading the circuit, reduce the coupling until further reduction does not change the susceptance variation ΔB_1. The short-circuit input susceptance is then given by the relation

$$B_{11} = - B_1, \qquad (7)$$

and the short-circuit input conductance by the relation

$$y_{11} = \frac{\Delta B_1}{2}\left[(1 + 2\eta^2)^{1/2} + \eta\right] - G_1. \qquad (8)$$

In most systems the inequality

$$\eta^2 = \left(\frac{\Delta G_1}{\Delta B_1}\right)^2 \ll 1 \qquad (9)$$

holds; thus (8) may be approximated by the equation

$$y_{11} = \frac{\Delta B_1}{2}\left[1 + \eta + \eta^2\right] - G_1 \qquad (10)$$

or even further by the relation

$$y_{11} = \frac{\Delta B_1}{2} - G_1 \qquad (11)$$

if η is negligible.

An alternative method in which the signal frequency remains constant during the test is based upon the circuit shown schematically in Fig. 65. For y_{11} the measurement is carried out as follows: the oscillator is coupled to a tuned circuit through a cutoff attenuator that has a very high resistance (a waveguide operated far below its cutoff frequency). The voltage across the tuned circuit is amplified by an amplifier and measured by voltmeter V. Before the tube is inserted, the circuit is tuned for resonance as indicated by a maximum reading of V, and the calibrated setting of the tuning capacitor in the tuned circuit is recorded. After the tube is inserted, the circuit is again tuned for resonance, and the calibrated capacitor is read again. The difference between this reading and the first reading is equal to ΔC. From ΔC the value of B can be calculated by means of the formula:

$$B = \omega \Delta C.$$

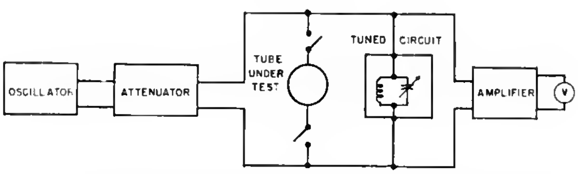

Fig. 65—Circuit arrangement for measurement of short-circuit input admittance y_{11} at constant frequency.

The reading of V is then brought to the original value by adjusting the attenuator. The value of the input resistance $R = (1/G)$ of the tube can be calculated from the relation

$$R = \frac{R_T}{10^{P/20} - 1},$$

in which R_T is the resistance of the tuned circuit and P is the difference in decibels between the two positions of the attenuator.

7.3.1.2 Measurement of y_{12} (feedback transfer admittance):

a) With the input termination still set at the value for resonance obtained in Step c) above, excite the output circuit through Y_{C2}. In the event that oscillation difficulties are encountered, detune the output circuit Y_2 or load it until oscillation stops.

b) Record the voltmeter readings V_1 and V_2.

The magnitude of the feedback transfer admittance is then given by the relation

$$|y_{12}| = \left|\frac{V_1}{V_2}\right| \frac{\Delta B_1}{2}\left[(1 + 2\eta^2)^{1/2} + \eta\right], \qquad (12)$$

and if

$$\eta^2 = \left(\frac{\Delta G_1}{\Delta B_1}\right)^2 \ll 1,$$

(12) may be simplified to

$$|y_{12}| = \left|\frac{V_1}{V_2}\right| \frac{\Delta B_1}{2}\left[1 + \eta + \eta^2\right] \qquad (13)$$

or

$$|y_{12}| = \frac{\Delta B_1}{2}\left|\frac{V_1}{V_2}\right|, \qquad (14)$$

where ΔG_1 and ΔB_1 are the values obtained in the preceding measurement of y_{11}.

7.3.1.3 Measurement of y_{22} (short-circuit output admittance): The short-circuit admittance may be measured by following the procedure outlined for the input coefficient y_{11}, the signal being coupled through Y_{C2}. If the subscripts 1 and 2 are interchanged, all of the foregoing formulas concerning y_{11} may be used to relate y_{22} to the measured data.

7.3.1.4 Measurement of the magnitude of y_{21} (forward transfer admittance): The magnitude of the forward transfer admittance may be measured by following the procedure outlined previously for the measurement of the magnitude of y_{12}. If the subscripts 1 and 2 are interchanged, all of the formulas concerning y_{12} may be used to relate y_{21} to the measured data.

7.3.1.5 Susceptance-variation equipment: Fig. 66 is a cut-away assembly drawing of a susceptance-variation circuit in which 1 and 2 are adjustable-length wave coils[5] comprising the main parts of the terminations Y_1 and Y_2. Variable micrometer capacitors 3 and 4 provide a calibrated small variation in the susceptance of Y_1 and Y_2 when the coarser adjustments of the wave coils are too large to give the precision necessary to measure small conductance components of the admittances. Adjustable coupling capacitors 5 and 6 are illustrated schematically as Y_{C1} and Y_{C2} in Fig. 64. Adjustable

[5] A wave coil is a coaxial line having a coiled center conductor.

probes 7 and 8 provide for coupling samples of the voltages V_1 and V_2 to a heterodyne receiver. These probes may be replaced by crystal voltmeters.

If the conductance being measured is small enough to be measured by the available range of the micrometer capacitor, the susceptance and conductance variations between $1/\sqrt{2}$ voltage points are given by

$$\Delta B = \omega \Delta C$$

$$\Delta G = 0,$$

where ω is the angular frequency of measurement and ΔC is the micrometer capacitance variation between the half-power points. If the conductance to be measured is too large to be measured by the micrometer variation, the susceptance variation may be obtained by varying the wave-coil length. The quantities ΔB and ΔG may then be obtained directly from the calibration of the termination-circuit conductance and susceptance.

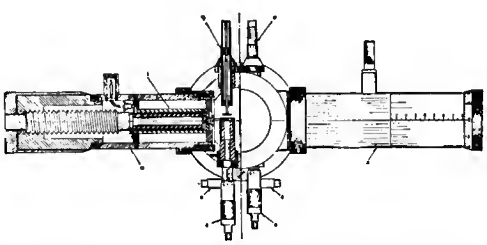

Fig. 66—Cut-away assembly drawing of a susceptance-variation circuit using wave coils.

The calibrations are obtained by terminating the wave coil in various standard capacitors at a number of frequencies such that, for any one frequency, the wave coil can be resonated at a number of lengths. These small capacitors are made in such a form that the conductance is negligible and the capacitance is the LF value over the entire frequency range. Consequently, the conductance of the circuit is

$$G_1 = \frac{\omega \Delta C}{2}, \tag{16}$$

and the susceptance is

$$B_1 = -\omega C_s,$$

where ω is the resonant angular frequency and ΔC is the total capacitance variation between half-power points when the circuit is terminated in a standard capacitance C_s.

7.3.1.6 Coaxial-line termination admittances: To extend the upper frequency limit of the susceptance-variation set described above, the adjustable wave coils may be replaced by adjustable-length coaxial lines, as shown in Fig. 67. By making the maximum electrical length of these lines slightly greater than that of the wave coils, a comfortable overlap in frequency can be obtained. Line-type terminations may be calibrated in the same manner as wave coils, by means of similar standard capacitors. The measurement operations proceed in a manner similar to that described for the wave-coil setup.

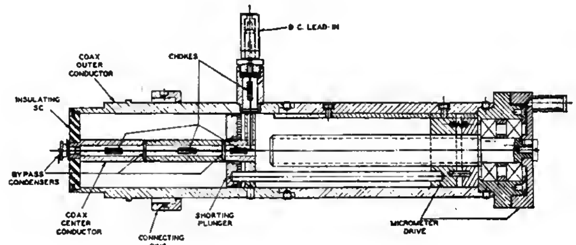

Fig. 67—Cut-away assembly drawing of a susceptance-variation circuit using coaxial lines.

7.3.2 Resistance-Substitution Method of Measurement: The resistance-substitution method applies only to the measurement of the conductive component of the admittance. The susceptive component must be measured by means of a calibrated susceptance element, as described in Section 7.3.1. Ideally, the resistance-substitution method involves, in the case of a two-terminal admittance, the removal of the electron-tube transducer from the calibrated admittance element and its replacement by a standard pure resistance of such a value that the voltage reading between the two terminals is the same as that obtained with the transducer in place. (See Fig. 68.) The calibrated susceptance element must be adjusted to resonance both before and after substitution of the resistance. If the measurement is of y_{11}, the following relations are obtained:

$$B_{11} = -B_1, \tag{17}$$

as in Section 7.3.1 above, and

$$G_{11} = 1/R, \tag{18}$$

where R is the standard resistance of a value that satisfied (18).

There are practical difficulties in obtaining standard resistors having negligible reactance at frequencies of the order of 10^8 cps or higher. Wire-wound resistors are not usable at such frequencies. The most satisfactory types available are the metalized-glass or ceramic-rod resistors of relatively small physical size, having low-inductance terminals and very little distributed capacitance. A further difficulty arises from the fact that such resistors are obtainable only in discrete values of resistance. It would not be practicable to obtain the very large number of resistors needed to match the resistance of any electron-tube transducer. Hence, it is necessary to utilize a transformation property of the admittance-measuring equipment in order to match any arbitrary admittance with some one of a reasonably small set of standard resistors. A suitable resistance-substitution set consists of a transmission line of length l short-circuited at one end, having a characteristic admittance Y_0 and a propagation constant γ. If a known admittance

Y is placed across the line at a distance x from the short-circuited end, as shown in Fig. 69, the admittance Y_t at the open end of the line is given by the relation

$$Y_t = Y\left[\frac{\sinh \gamma x}{\sinh \gamma l}\right]^2\left[\frac{1}{1+\dfrac{Y}{Y_0}\dfrac{\sinh \gamma x}{\sinh \gamma l}\sinh \gamma(l-x)}\right]. \quad (19)$$

Conversely, the admittance Y_t is the admittance that would have to be placed at the open end of the line to produce the same effect there as the known admittance Y at the position x. Eq. (19) then represents the property of the transmission line of converting admittance Y at position x into admittance Y_t at position l. This expression can be simplified for a low-loss line having a characteristic admittance Y_0 large compared with the bridging admittance Y. Thus, if $Y_0 \gg Y$,

$$\left|\frac{Y}{Y_0}\frac{\sinh \gamma x}{\sinh \gamma l}\sinh \gamma(l-x)\right| \ll 1, \quad (20)$$

and the real part of the propagation constant γ of the line is small,

$$\left[\frac{\sinh \gamma x}{\sinh \gamma l}\right]^2 = \left[\frac{\sinh \beta x}{\sinh \beta l}\right]^2, \quad (21)$$

where $\beta = 2\pi/\lambda$. Eq. (19) then simplifies to

$$Y_t = Y\left[\frac{\sin \beta x}{\sin \beta l}\right]^2. \quad (22)$$

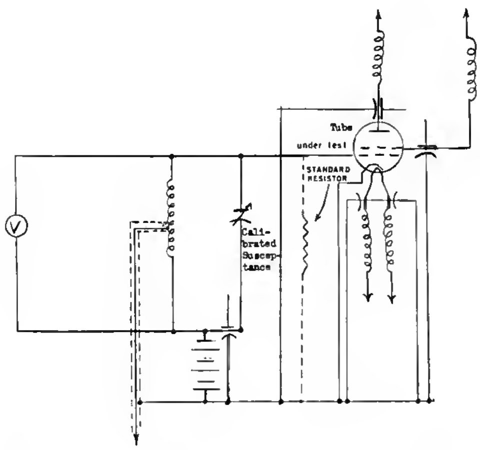

Fig. 68—Circuit arrangement for measurement of admittance by the resistance-substitution method.

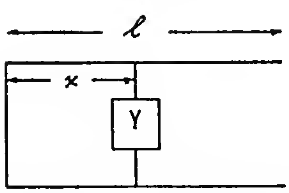

Fig. 69—Transmission-line admittance transformer.

If Y is a pure conductance of value $1/R$, then Y_t is the pure conductance

$$G_T = \frac{1}{R}\left[\frac{\sin \beta x}{\sin \beta l}\right]^2. \quad (23)$$

In the measurement of the short-circuit input admittance y_{11}, a low-loss transmission line of large characteristic admittance is coupled loosely to an oscillator near the short-circuited end. At the open end are a voltage-detecting device and a calibrated capacitor, by means of which β_{11} is obtained from (17). With the unknown transducer across the open end and the capacitor adjusted to obtain resonance, a voltage reading is taken. The transducer is removed and one of the standard resistors placed across the line bridging the two conductors. The position of this resistor along the line is then adjusted and the system is readjusted for resonance with the calibrated capacitor until the voltage, as measured at the end of the line, is the same as before. By (23) we have

$$G_{11} = G_T\frac{1}{R}\left[\frac{\sin \beta x}{\sin \beta l}\right]^2. \quad (24)$$

where R is the resistance of the standard placed x cm from the short-circuited end of the line. It is evident that a resistor must be selected having a resistance value near to but not larger than the reciprocal of G_{11}.

Since the transmission line should have low-loss and low-characteristic impedance, a coaxial line is desirable. The line will require a longitudinal opening or slot in order to permit one of the standard resistors to bridge the line at an adjustable position to satisfy the required voltage condition. An electron-tube voltmeter is capacitively coupled to the open end of the line across which the electron-tube transducer may be attached. Socket and filter arrangements for wire-lead tubes can be attached to this line. Radiation difficulties arising from the longitudinal opening in the line, together with the increasing difficulty in obtaining resistance standards at frequencies much above 3×10^8 cps appear to make this type of measuring equipment impracticable for measurements at higher frequencies on surface-lead tubes.

8. Nonlinear Characteristics

8.1 Detection Characteristics

The following characteristics are of interest in connection with large-signal detection.

8.1.1 Rectification Characteristic: In the general case of a tube of n electrodes, the connections are shown in Fig. 70. E is an ac generator considered as having zero dc and ac impedance. All electrodes not entering directly into the measurements are maintained at steady and specific voltages.

The average currents in an electrode circuit, as read by a dc instrument, are plotted as ordinates against values of the direct voltage E_j on the electrode as

abscissas for various values of E as a parameter; *i.e.*, E is held constant for each graph.

8.1.2 Transrectification Characteristic: The transrectification characteristic is the graph between the average current in the circuit of an electrode, the direct voltage on that electrode, and the amplitude (or root-mean-square value) of an alternating voltage impressed on another electrode. The connections for this test for a tube of n electrodes are shown in Fig. 71. The electrode j and other electrodes are to be maintained at their specified values of direct voltage.

The values of direct voltage E_k in the electrode circuit k are plotted as abscissas against the average of current I_k in that circuit as ordinates for various values of alternating voltage E applied to the other electrode as a parameter; *i.e.*, E is held constant for each graph.

8.2 Conductance for Rectification

Conductance for rectification is most simply determined from the slope of the graph showing the relation between the values of the average direct currents in the circuit of an electrode as ordinates, and the direct voltages in the circuit of the same electrode as abscissas, with a constant specified RF voltage applied to one or more of the electrodes.

A balance method for measuring conductance for rectification is also available. An application of this method to the measurement of the plate conductance for transrectification in a triode is shown in Fig. 72. In this case the voltage is applied to the grid.

$$g_p' = \frac{R_1}{R_2 R_3}.$$

The plate resistance for rectification r_p' is the reciprocal of the plate conductance for rectification; *i.e.*,

$$r_p' = \frac{1}{g_p'} = \frac{R_2 R_3}{R_1}.$$

In Fig. 72 capacitor C is a RF by-pass. C_n is necessary to balance the tube capacitance and the capacitance of C. The resistive elements of the bridge are balanced in the usual manner.

Although Fig. 72 shows only the measurement of the plate conductance for transrectification of a triode, the method is also applicable to the measurement for conductance of any electrode for ordinary rectification or transrectification. For multielectrode tubes, all electrodes not directly involved in the measurement should be maintained at constant and specified voltages.

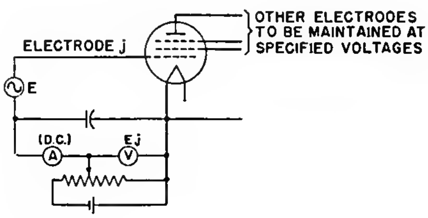

Fig. 70—Circuit arrangement for measuring rectification characteristic.

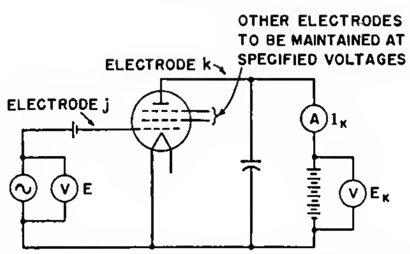

Fig. 71—Circuit arrangement for measuring transrectification characteristic.

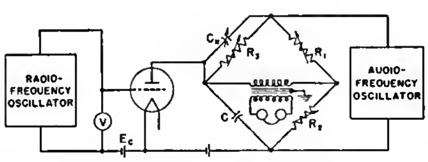

Fig. 72—Circuit arrangement for measuring (plate) conductance for rectification.

9. Audio Power Output

The power output of a vacuum tube is dependent upon the direct operating voltages applied to the various electrodes, the external load impedance in the plate circuit, and the magnitude of the exciting voltage applied to the control grid. In any case, the operating conditions are subject to the maximum safe values placed by the manufacturer upon electrode voltages, electrode power dissipation, and space current drawn from the cathode.

For vacuum tubes normally used as class-A amplifiers under conditions such that the control grid is not driven appreciably positive with respect to the cathode, the power output is the power delivered to a resistive load with a sinusoidal input voltage applied to the grid. For tubes in which the control grid is driven positive, as in class-B amplifier tubes as usually operated, special consideration must be given to the impedance in the grid circuit and its effect on harmonic distortion. Further consideration is given to amplifiers of this class in Section 9.2.

In general, where harmonic distortion is undesirable, the power output available in any particular application will increase with the permissible percentage of harmonics. The amount of distortion that may be tolerated varies greatly in different applications; consequently, no single criterion of permissible distortion is acceptable in all cases. A reference to the available power output for sinusoidal input should be accompanied by a statement of the maximum percentage of distortion present at this power output or at lower values of power output within the operating range. This percentage of distortion is expressed in terms of the total distortion as defined in Section 9.1 below, or the individual harmonic components of output current

may be expressed separately as percentages of the current of fundamental frequency. Usually, the second and third harmonics will suffice, but higher-order terms should be given where they are of the same order of importance as the second and third harmonics.

9.1 *Measurement of Harmonics*

The total harmonic distortion is expressed by the relation

$$D = \frac{(I_2^2 + I_3^2 + \cdots + I_n^2)^{1/2}}{I_1},$$

where I_1 is the amplitude of the fundamental, and I_2, I_3, $\cdots$, I_n are the amplitudes of the 2nd, 3rd, $\cdots$, nth harmonics of the current in the load.

The distortion may be measured by a harmonic analyzer, of which several types have been described in the literature. When merely the value of D is desired, those analyzers which measure the root-mean-square value of all harmonics present are preferable to those which measure the separate harmonics.

The method of Suits[6] is a particularly good example of the type of analyzer which measures the harmonics separately. The Suits method requires only the simplest apparatus, and where laboratory facilities are limited this advantage may outweigh the disadvantages involved in the computation of D.

The Belfils analyzer[7,8] utilizes an ac Wheatstone-bridge balance for the suppression of the fundamentals, and is particularly useful for the direct measurement of D. For maximum convenience, the frequency of the audio-frequency source should be very stable. This instrument can be operated so that it is direct reading by maintaining a constant input voltage.

In the McCurdy-Blye analyzer[9] low- and high-pass filters are used to separate the harmonics from the fundamental. This instrument is superior to the Belfils type in that the frequency of the source may vary somewhat without necessitating readjustment.

9.1.1 Precautions: The sinusoidal voltage applied to the control grid should be free from harmonics. This can be assured by the use of a filter (see Fig. 73).

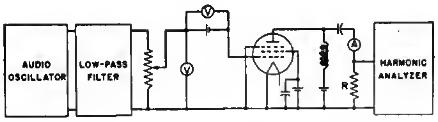

Fig. 73—Circuit arrangement for measuring undistorted power output of a pentode.

[6] C. G. Suits, "A thermionic voltmeter method for the harmonic analysis of electrical waves," Proc. IRE, vol. 18, pp. 178–192; January, 1930.

[7] G. Belfils, "Mesur du 'residu' des courbes de tension par la methode du pont filtrant," *Rev. Gen. d'Elect.*, vol. 19, pp. 526–529; April, 1926.

[8] I. Wolff, "The alternating current bridge as a harmonic analyzer," *J. Opt. Soc. Am.*, vol. 15, pp. 163–170; September, 1927.

[9] R. G. McCurdy and P. W. Blye, "Electrical wave analyzers for power and telephone systems," *J. AIEE*, vol. 48, pp. 461–464; June, 1929.

If an iron-cored choke is employed for shunt feed in the plate circuit (Fig. 73), care should be exercised in its selection or design to avoid the generation of harmonics in it as a result of the nonlinear and hysteretic behavior of the iron.

9.2 *Measurement of Power Output at Audio Frequency*

In the measurement of power output use is made of well-established measurement technique. However, a number of general precautions should be observed and are given below.

9.2.1 Precautions: In class-A_1 amplification the grid is not driven positive with respect to the cathode; hence, the peak grid input voltage will be approximately equal to grid bias.

The condition that no appreciable current shall flow in the grid circuit may require the peak grid-input voltage to be slightly less than the grid bias, especially in filamentary tubes whose filaments are heated by alternating current.

When the grid is driven positive, the essential characteristics of the driving circuit should be specified.

The effects of the regulation of the power-supply voltages should be taken into consideration.

The effects of feedback due to common circuit elements should be considered.

9.3 *Measurement of Push-Pull Power Output at Audio Frequencies*

The techniques used in the measurement of power output in push-pull operation are basically similar to those employed for single-ended operation, but some additional precautions should be observed and are given below.

9.3.1 Precautions: The input signal should be substantially free from distortion.

Care should be taken to insure that the input transformer produces voltages that are equal, opposite in phase, and without harmonics. These requirements are most readily met if the transformer is of adequate size for the power required and has low leakage inductance between windings. When the grids are driven positive, as in Class AB_2 or Class B operation, the effective resistance and impedance of the driving source must be specified. The effects of regulation of the power-supply voltages, including grid bias, are particularly important in Class AB_2 and Class B operation.

If a center-tapped iron-cored choke is employed for shunt-feeding the anode circuit, it should be designed so that the two halves of the winding have the same dc resistance, the leakage inductance between the two halves of the winding is low, and the iron core is of adequate size and correct material to avoid the generation of harmonics.

In reporting measurements of push-pull stages using self-bias it should be stated whether the cathode resistance was by-passed.

10. Radio-Frequency Operating Tests for Power-Output Tubes

10.1 Determination of RF Power Output

The test for power output is made by operating the tube as a RF oscillator or as a RF amplifier. The power output can be determined either 1) directly by measuring the RF power delivered to a load and correcting for circuit losses or, 2) indirectly from the difference between the total input power and the power dissipation in the tube.

10.1.1 Direct Measurement of RF Power Output: Power output may be measured by a number of methods, the most suitable being dependent upon the magnitude of power to be measured and the frequency range involved in any particular application. All methods involve a device designed to absorb the power to be measured as well as a device for monitoring that power. At low output levels the monitoring device may be capable of absorbing the total power developed.

Some suitable devices for the absorption of power are 1) load lamps, 2) resistors, either radiation- or convection-cooled, 3) dissipative transmission lines or waveguides, or 4) radiating antennas, either in air or in a dissipative medium.

Suitable monitoring equipment include 1) devices for measurement of light intensity, such as photometers; 2) devices for the measurement of temperature, such as thermocouples, thermometers, and thermistors; and 3) voltage-, current-, and power-measuring devices.

10.1.2 Load-Lamp Method: The power to be measured is dissipated in a lamp load across a coil suitably coupled to the output circuit and tuned to optimum power absorption by means of a series or parallel capacitor, as shown in Fig. 74. The lamp is calibrated at a low frequency or on dc. Instead of the resistance bridge shown as a monitor, a light-measuring device may be used.

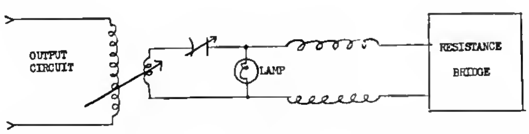

Fig. 74—Circuit arrangement for measurement of RF power output with a load-lamp.

10.1.2.1 Precautions: At higher frequencies, it may be necessary to remove the base from the load lamp to avoid losses. To improve the consistency of the results, it may be necessary to age the load lamp for several hours and to chose a load lamp that will not be up to full brilliancy at the power levels to be measured.

10.1.3 RF Absorption-Wattmeter Method: The RF absorption wattmeter comprises a resistor of specified resistance that provides an RF load of low SWR. The power dissipated in the resistor is measured by a tube (or crystal-diode) voltmeter bridged across the input. The voltmeter may be calibrated in watts and may provide more than one range. At high powers, the resistor may require special cooling by forced air or circulating water.

10.1.4 Indirect Measurement of RF Power: When it is possible to obtain accurate measurement of plate dissipation by the methods given in **11.1**, the power output may be determined by subtracting the dissipation from the total power input to the plate circuit. Approximate corrections should be made in an amplifier circuit for the power obtained from the driving source.

10.2 Grid Driving Power

The grid driving power or excitation power of an amplifier tube may be measured by a substitution method in which a calibrated load is substituted for the tube. The load is adjusted so that the same conditions are obtained in the driving circuit as with the amplifier tube connected. The power measured in this load will then be equivalent to the grid driving power of the amplifier tube.

A general method suitable for all frequencies up to approximately 200 Mc employs the circuit of Fig. 75. The following procedure is used: The normal grid-tuned circuit is represented by L_1, C_1 and the substituted load is represented by C_2 and R_L, where R_L is a lamp capable of dissipating about twice the grid driving power of the tube under test. The combination of diode D, resistance R, and microammeter M serves as a peak voltmeter to measure the grid-cathode RF voltage.

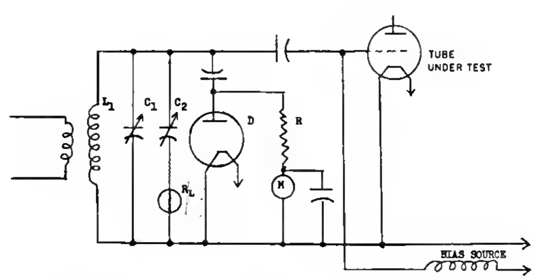

Fig. 75—Circuit arrangement for measuring grid driving power up to about 200 Mc.

The tube under test is set up under normal conditions, but capacitor C_2 is adjusted so that a measurable output is obtained in lamp R_L. This power is measured by any standard comparison method, and the reading of M is noted. The test tube is then removed and the grid circuit tuned to resonance by adjusting C_1, resonance being indicated by a maximum deflection of M. The capacitance C_2 is then increased, C_1 being simultaneously adjusted to keep the grid circuit at resonance until the same reading is obtained on M as under normal conditions. The power in lamp R_L is again measured. The change of power is a measure of grid driving power.

Another method suitable for measuring the driving power of an amplifier tube at frequencies above approximately 200 Mc is the use of a transmission line or wave-

guide as a power-measuring device. It is convenient to use a section of transmission line not shorter than one-half wavelength inserted between the driving source and the amplifier tube. If the maximum and minimum rms voltages V_{max} and V_{min}, respectively, are measured along a low-loss line, the power supplied to the amplifier is

$$\text{Driving Power} = \frac{V_{max} \times V_{min}}{Z_0},$$

where Z_0 is the characteristic impedance of the line in ohms. The line must be sufficiently well matched so that V_{min} can be measured with the desired accuracy.

The following methods generally apply only to measurements at frequencies below about 10 Mc.

A method applicable to either an oscillator or an amplifier is the use of an oscillograph to measure the grid current and voltage and their phase relation at the input terminals of the tube. The average power can then be determined by graphical integration.

Grid driving power can also be determined for either an oscillator or an amplifier tube by measuring the direct component of grid current and the peak value of the grid driving voltage. The product of these quantities gives an approximation of grid driving power if the dielectric and lead losses and the effects of transit time are negligible.

The power delivered to the input of the tube can be determined by subtracting the product of the direct grid current and the bias voltage from the grid driving power.

11. Electrode Dissipation and Bulb Temperature

Three types of cooling are generally employed for vacuum-tube electrodes. These are radiation, liquid, and forced-air cooling. Methods for measuring power loss are different for each type.

11.1 Methods of Measuring Anode Dissipation

11.1.1 Radiation-Cooled Anodes:

11.1.1.1 Optical-pyrometer method: This method is applicable only to radiation cooling where the anode is radiating in the visible spectrum under oscillating conditions. A pyrometer is used to measure plate temperature at the hottest point and at two other points having different temperatures. The tube is then operated statically with a relatively low direct-anode potential, and with adequate alternating or positive direct grid voltage to reproduce as well as possible the dissipation pattern of the previous oscillating conditions. When the pyrometer readings at the chosen three points match the earlier data, the dc input to the anode gives a fair measure of the anode dissipation under the oscillating condition. This method can be accurate to within a few per cent.

11.1.1.2 Thermocouple method: The anode radiation may be collected from one side of the anode at a time by means of a conical tubing with reflecting walls, and the energy focused upon a thermocouple. The anode is then supplied with dc power under static conditions to duplicate the previous thermocouple readings, and the dc power supplied will be equivalent to the anode dissipation.

11.1.1.3 Calorimeter method: The total power lost in the tube, including filament and grid loss, can be measured by immersing the tube in a circulated liquid and measuring the rate of flow and temperature rise of the liquid after the temperatures are stabilized. The power dissipated can then be calculated from the formula given in Section 11.1.2. In this method, also, static inputs that result in static dissipation only may be used to match the dynamic conditions.

11.1.2 Liquid-Cooled Anodes: The direct method of measuring anode dissipation of liquid-cooled anodes consists of measuring the flow of liquid through the cooling jacket and the temperature rise between inlet and outlet points. The total dissipation is proportional to the product of the rate of flow and the temperature difference. Since the filament or heater power alone may also be measured by this method, the net anode dissipation can be calculated. In this method, also, measured static inputs that result in static dissipations only may be used to match the dynamic conditions.

11.1.3 Forced-Air-Cooled Anodes: For forced-air cooling, quite satisfactory results may be obtained by a temperature-matching method. A temperature-responsive device is placed at some point in the cooling system. With a constant flow of air at constant inlet temperature, the indication of this device is noted under oscillating conditions. DC power is then supplied to the anode under nonoscillating conditions to duplicate this indication. This power is equal to the anode dissipation under oscillating conditions. Various temperature-responsive devices may be used, provided they are not affected by RF fields.

11.1.3.1 Cooler-temperature matching method: In this method the temperature-indicating device is brought into intimate contact with the metal cooler attached to the anode. Under dc conditions, the anode voltage should be relatively low and the grid voltage positive in order to produce uniform heating.

11.1.3.2 Air-temperature matching method: In this method the temperature-indicating device is placed in the outgoing air stream, which is preferably confined within a duct to prevent disturbing air currents. An insulated duct diverting a part of the outgoing air may also be used for this purpose. The indicating device may be, for example, one or more thermomemters, one or more thermocouples connected in series, or a resistance-wire grid. With the latter, measurement of resistance gives an indication of temperature.

11.2 Methods of Measuring Grid Dissipation

Information on grid dissipation in power tubes is important to insure that the grid is operated below the

point at which primary emission becomes excessive. The methods of measurement given here are applicable only at frequencies low enough so that electron transit time is negligible. It should be borne in mind that grid dissipation is a component of, but not the same as, grid driving power.

11.2.1 Direct-Grid-Current Method: The grid dissipation can be calculated if the peak alternating grid voltage is known. The product of the direct grid current and the peak positive excursion of the grid voltage with respect to the cathode is the approximate grid power dissipation. This method is subject to error if secondary emission is appreciable.

11.2.2 Graphical Integration Method: The grid dissipation can be calculated from the dynamic characteristic of the tube by graphical integration. This method also is subject to error if secondary emission is appreciable.

11.2.3 Liquid-Cooled Grids: In tubes with liquid-cooled grids the rate of flow and the temperature rise of the cooling medium will give the grid power loss under oscillating conditions by the method described under Section 11.1.2. Subtraction of the grid loss due to the filament alone gives the grid dissipation caused by oscillation. Here, also, measured static inputs which result in static dissipation only may be used to match the dynamic conditions.

11.3 Methods of Measuring Bulb Temperature

There are several alternative methods which differ in accuracy and convenience of use in particular circumstances. Tubes may be rated for 1) the temperature at the hottest point on the envelope, or 2) the maximum average circumferential temperature.

11.3.1 Temperature at a Point: The method for measuring the point of highest temperature on the bulb employs a thermocouple. The possibility that high temperature gradients may exist should be recognized in probing for the point of highest temperature. To obtain the greatest accuracy, the thermocouple must be constructed of wires of the order of 0.004-in diameter, but not greater than 0.005-in, and the wires adjacent to the junction must be held in good thermal contact with the bulb to prevent the wires from cooling the junction by conduction and radiation.

11.3.2 Average Circumferential Temperature: Thermocouple wires are welded to two diametrically opposite points on a close fitting phosphor-bronze split ring. Commonly the ring has a 0.1-in$\times$0.02-in cross section. It must fit tightly around the bulb. Depending upon design and ambient conditions, this method yields temperature readings approximately 10–20°C lower than the method described in 11.3.1.

11.3.3: Where the above methods cannot be used (for example, where the bulb is surrounded by a shield), a less accurate measurement can be made with temperature indicating paints. Care and judgement should be exercised in their use.

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Part 2

Cathode-Ray Tubes

Subcommittee 7.2
Cathode-Ray Tubes

A. M. Morrell, *Chairman* (1961–1962)
G. W. Pratt, *Past Chairman* (1958–1960)
J. R. Adams, *Past Chairman* (1956–1957)
C. G. Lob, *Past Chairman* (1955)

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H. J. Evans	R. S. Hunter	W. B. Nottingham
J. P. Foltz	J. T. Jans	E. W. Shinholt
H. K. Hammond III	L. T. Jansen	D. D. Van Ormer
	G. P. Kirkpatrick	

1. Introduction

1.1 Scope

This part describes the method of measurement of important characteristics of those types of cathode-ray tubes classified as oscilloscope tubes, monochrome picture tubes, and color picture tubes. Methods of test used primarily for quality control of the products during manufacture are not included.

1.2 Reference to Methods of Testing Other Tubes

A number of the general electron-tube precautions and test conditions of Part 1, Introduction, are applicable to cathode-ray tubes and should be observed in addition to those given in this part.

1.3 Precautions

1.3.1 Magnetic Shielding: Stray magnetic fields through a cathode-ray tube during test can deflect and distort the beam sufficiently to give misleading test results. Fields through the tube near its screen will usually only shift the spot position, while fields through the electron-gun assembly can distort or even cut off the beam by decentering it in the focusing fields and in the limiting apertures. It is therefore desirable to provide magnetic shielding for a cathode-ray tube under test to reduce or eliminate the effect of undesired constant or varying magnetic fields.

Shields are preferably made of a high-permeability alloy and should be of such design as to minimize undesired magnetic fields. In certain tests the earth's magnetic field, or a field to simulate the average earth's field, may be used in testing so that the conditions of test may represent actual conditions in unshielded operations.

1.3.2 Demagnetization of Parts: Tube parts made of ferromagnetic materials can become magnetized and produce anomalous effects. As a general rule, therefore, tubes containing ferromagnetic materials must be demagnetized prior to testing.

1.3.3 Insulation: In testing cathode-ray tubes insulation and spacing must be provided in the test equipment to prevent arcing and leakage that may change the operating conditions imposed on the tube or give false current readings. In general, cathode-ray tubes are operated so that those electrodes upon which the highest-frequency signals are impressed are at or near ground potential. Monochrome picture tubes and color picture tubes are commonly operated with the cathode or control grid at ground potential, and oscilloscope tubes are generally operated with the deflecting electrodes near ground potential.

Leakage currents and charges on the external surface of the tube can produce spot shifts and distortion of the screen patterns. One method of reducing these effects is to test the tube with the screen near ground potential.

1.3.4 Safety Precautions: In testing cathode-ray tubes care should be taken to insure that the operator is protected against high-voltage shock, X-ray exposure, and tube implosion.

1.4 Ambient Light

Several cathode-ray tube tests depend upon measurement of light output and color from the tube screen, or on visual observations of the fluorescent or phosphorescent screen patterns. It is essential that the effect of ambient light be kept negligible in comparison with the screen luminance during these tests in order to avoid errors in measurements.

1.5 Operating Conditions

1.5.1 Applied Voltages: The various voltages must be applied in such sequence as to prevent tube damage.

In applying the voltages care must be taken to insure that the maximum rated voltages between electrodes are not exceeded. The control electrode voltage must always be of such value as to prevent screen burning during testing.

1.5.2 Regulation: High-voltage power supplies are often designed with poor regulation. If the regulation of the power supply used is such that any changes occur in the electrode voltages as a result of changes in electrode currents, the electrode voltages must be readjusted to the desired values.

1.5.3 Filtering: Undesired alternating components in the source voltages can result in errors of measurement. The magnitude of these errors depends upon the relative magnitude of the alternating and direct components of the source voltages.

1.5.4 Focus: Improper focus adjustment of cathode-ray tubes may result in misleading data on such items as luminance, resolution, and chromaticity. The focusing field must therefore be adjusted for the desired focus, depending upon the type of measurement being made.

1.5.5 External Magnetic Components: Many tubes require the use of one or more external magnetic components. These components include either permanent magnets or electromagnets or both. The strength of the field of these magnets is an important part of their characteristics and may be altered by sharp blows, heat, or the presence of other strong magnetic fields. The number and type of external magnetic components used during testing must be as specified for the particular tube type.

The precise location and orientation of these external components with respect to the tube is of the greatest importance in obtaining reproducible test results. In general, the external components must be located as required by the specification for the tube type under test. The location of the external magnetic components can affect such tube parameters as focus, luminance, current distribution among electrodes, color uniformity, convergence, raster geometry, etc. Diverse tests require different positions and/or field strengths of the components.

The distribution and uniformity of the field produced by these components have an important effect on the performance of the tube. In comparing the performance of tubes it is important to know the complete characteristics of these external components.

Some examples of external magnetic components are: focusing coils or focusing magnets, magnetic deflection yokes, ion-trap magnets, centering devices, beam-aligning magnets, convergence magnets, and color-purity magnets.

1.5.6 Deflecting-Electrode Potentials: The zero signal potential of the deflecting electrodes must, unless otherwise specified, be equal to the potential of the electrode through which the electrons pass just before entering the deflecting field. With tubes designed for balanced deflection, care must be taken to insure that the deflecting signals are balanced in order to minimize distortion.

2. Instructions for Test

For general instructions refer to Part 1, Introduction.

2.1 Cutoff Voltage

2.1.1 Luminance-Cutoff Voltage: The luminance-cutoff voltage is determined by measuring the control-electrode bias voltage for visual extinction of the undeflected focused spot or the scanned raster. Since these values will in general be different, the method of determination of the cutoff voltage (*i.e.*, spot or raster) must therefore be specified. The ambient illumination at the tube screen must be kept at such a low value that further reduction will not affect the measured value. In measuring spot cutoff the viewing distance, the use of optical instruments for viewing the spot, or the extent of dark adaptation of the observer may affect the measured value.

2.1.2 Current-Cutoff Voltage: The current-cutoff voltage is determined by measuring the control-electrode bias voltage required to reduce the cathode current to some specified low value, usually about 1 μa or less.

2.2 Leakage Currents

Leakage currents are the undesired currents which flow between the electrodes when voltages are applied to those electrodes with the electron beam cut off. Most leakage currents can be conveniently determined by applying all the specified electrode voltages and reading the currents in external circuits with the control electrode bias set at a specified value beyond cutoff. These currents are identified by the circuits in which they are read.

If the tube application normally required substantial resistance in external circuits, it is desirable in the measurement of tube leakages to include an appropriate amount of resistance in the circuit, since measured values of the leakage without added circuit resistance may not correlate with performance in the intended application.

In the case of electrostatic-deflection tubes the measurement of deflecting-electrode leakage currents is difficult to make if the deflecting electrodes are substantially above ground potential. It is then convenient to measure the effective deflecting-electrode leakage in the following manner. The cathode-ray tube is operated at the specified electrode potentials with an undeflected focused spot on the screen. The control-grid potential is adjusted to limit the beam current to a value low enough to avoid burning of the screen (usually less than 1 μa). Each deflecting electrode is connected to the source of reference potential through a specified resistance. The value of this resistance must be the same for each deflecting electrode. The resistance in each deflecting-electrode circuit is short-circuited in turn, and the displacement of the undeflected focused spot is observed.

The amount of this displacement is a measure of the leakage in each deflecting-electrode circuit.

2.3 Electrode Currents

The current in the external circuit of each electrode is measured. Each current is identified by the circuit in which it is measured and must be corrected for leakage. It is frequently not possible to measure the current from certain individual electrodes, particularly screen current, since several electrodes may be electrically connected within the tube.

Precaution: When making certain electrode current measurements, in particular, the zero-bias cathode current, care must be taken to prevent burning of the screen or damage to limiting apertures as the result of excessive heating. The screen may usually be protected by using a full raster scan or defocusing the tube during measurement.

2.4 Gas Content

Method A: The gas content can best be determined by making use of certain electrodes in the tube as a modified ionization gauge (see Part 1, Paragraph 5.2.2) to determine the gas ratio.

Method B: The other makes use of the presence of a cross on the screen of a gassy tube. The gas-cross method is applicable only to tubes having deflecting electrodes.

2.4.1 Gas-Ratio Test: The *gas ratio* is given by the equation $G=(N-L)/P$, where G=*gas ratio*, N=ion current, L=leakage current in the electrode circuit used to collect ions, and P=electron current that produces the ions. This ratio is customarily expressed in microamperes per milliampere. The electron current is usually measured in the cathode circuit. The number-one grid is biased to produce a specified cathode current (usually at least several hundred microamperes for good sensitivity). The number-two grid is biased positively (usually approximately 200 volts). All electrodes beyond the number-two grid are usually tied together and biased negatively by a specified small amount (about 25 volts). The ionization current is measured in this circuit. No external magnetic components are used during this measurement. The leakage current is measured under the same conditions, with the exception that the tube is biased beyond cathode-current cutoff by either cathode or number-one grid bias. Under either bias condition, the ion-collecting electrodes must be the most negative electrodes in the tube.

Precaution: While it is possible to establish correlation between the gas-ratio measurement and the actual amount of gas in the tube, a single calibration is not applicable to electron guns of different designs.

2.5 Cathode-Ray-Tube Capacitances

2.5.1 General: The direct, interelectrode capacitances to be measured in a particular cathode-ray tube are usually listed in the individual data sheet. In general practice the following capacitances are the most important, because high frequencies are most likely to be applied to these elements:

1) On all types, the capcitance between control grid and all other elements tied together.
2) On types with heater and cathode not internally connected, the capacitance between cathode and all other elements tied together.
3) On electrostatically deflected types, the capacitances between deflecting electrodes, pairs of deflecting electrodes, and other tube elements.
4) On multigun and split-beam types, the capacitances of each section and additional capacitances between elements of the several sections.

The capacitance between the internal and external conductive bulb coatings is also usually specified.

Metal base sleeves and external conductive coatings should be grounded for all tests unless otherwise specified. For circuits and general precautions for capacitance measurements.

2.5.2 Measurement of Capacitance between Tube Coatings: Internal and external conductive coatings on portions of the bulb wall form a capcitor whose dielectric is the glass of the bulb. The high-frequency measurement of this capacitance is complicated by the fact that coatings have resistance. Because of this, the measured capacitance falls off at high frequencies and increases with increase in number of contact points to the coating. It is therefore desirable that data on capacitance of the tube coatings include the test frequency and the method of connecting to the coating.

2.6 Focusing-Electrode Voltage of Electrostatic-Focus Types

2.6.1 Focusing-Electrode Voltage at Low Screen Current: The focusing-electrode voltage for best focus of the spot is read with the spot undeflected and with the control-electrode bias voltage adjusted for a value of screen current low enough to avoid screen burning.

2.6.2 Focusing-Electrode Voltage at Recommended Operating Conditions: The focusing-electrode voltage for the best focus at the center of the raster is read with the recommended raster size scanned on the screen and with the control-electrode bias voltage adjusted for a given value of luminance or current.

2.7 Focusing-Coil Current of Magnetic-Focus Types

All measurements of focusing-coil current should be made with the focusing coil in the specified location (see Section 1.5.5).

2.7.1 Focusing-Coil Current at Low Screen Current: The focusing-coil current for best focus of the spot is read with the spot undeflected and with the control-electrode bias voltage adjusted for a value of screen current low enough to avoid screen burning.

2.7.2 Focusing-Coil Current at Recommended Operating Conditions: The focusing-coil current for the best

focus at the center of the raster is read with the recommended pattern size scanned on the screen and with the control-electrode bias voltage adjusted for a desired value of screen luminance or current.

2.8 Deflection Factor of Electrostatic-Deflection Types

Either direct or alternating voltages can be used in measuring the deflection factor of a pair of deflecting electrodes. The deflection factor is usually expressed in volts per inch. The difference in potential applied between the deflecting electrodes to produce a measured deflection of the spot is measured.

If alternating voltage is used, the peak-to-peak potential difference is measured by means of a peak-to-peak voltmeter or suitable calibrated oscilloscope. If the alternating voltage is sinusoidal, it is measured with an rms voltmeter, and the reading converted to peak-to-peak voltage. If the alternating voltage is not sinusoidal, the use of an oscilloscope is advisable. It is recommended that the frequency of the alternating voltage used be in the audio-frequency range. If high frequencies are used, the frequency should be kept below the value at which transit time and lead reactance begin to increase the deflection factor.

2.8.1 Deflection-Factor Uniformity—Electrostatic Deflection: With the tube operated as above, the deflection factor is measured for two or more specified points of deflection along each axis. The deflection-factor difference in per cent is determined from the formula

$$DFD = \frac{DF_1 - DF_2}{DF_1} \times 100,$$

where DFD is deflection factor difference, DF_1 is the deflection factor at some arbitrary amount of deflection 1, and DF_2 is the deflection factor at some second arbitrary amount of deflection 2. Since the deflection factor varies with the amount of deflection, a complete description of deflection-factor uniformity can only be obtained from a curve of deflection factor as a function of amount of deflection.

2.9 Deflection Factor of Magnetic-Deflection Types

Either direct or alternating current can be used in measuring deflection factor. The deflection factor is usually expressed in amperes per inch. The current through the deflection yoke to produce a measured deflection of the spot is measured.

The deflection factor is a function of the deflection yoke and its location on the tube; therefore, the deflection yoke and its location must be specified (see Section 1.5.5).

If alternating current is used, the peak-to-peak current is measured by means of a peak-to-peak current meter or suitable calibrated oscilloscope. If the current is not sinusoidal, the use of an oscilloscope is advisable. It is recommended that the frequency of the alternating current used be in the audio-frequency range. If high frequencies are used, the frequency must be kept below the value at which transit time and deflection-yoke resonance begin to change the deflection factor. If the tube is used normally at a particular frequency, this frequency is used for the measurement.

2.10 Screen Luminance

Screen luminance is measured by means of a suitable photometer.

2.10.1 Instrumentation:

2.10.1.1 Visual photometers: Visual photometers, when properly used by a trained observer, give accurate luminance measurements. However, they are subject to errors arising from:

1) Poor color match between screen luminous output and reference source.
2) Color-vision abnormalities of observers.
3) Observer fatigue.

Making measurements with visual photometers is tedious and fatiguing to the observer. Visual photometers are not therefore generally useful for routine production measurements of cathode-ray tube luminance.

The outstanding advantage of a visual photometer[1] for measurement of cathode-ray tubes is that it uses the human eye as the integrating device for the pulsating light output of the tube. Since the calibration standards for photometers are steady light sources, the interpretation of the standard source compared to the pulsating light source such as the cathode-ray tube is very important. A photometer that does not integrate the light in the same manner as the eye may give erroneous results. With the visual photometer this proper integration is automatic, since the eye is used as the integrating device.

2.10.1.2 Physical photometers:

2.10.1.2.1 Photovoltaic type: This type is probably the most widely used in industry because of its simplicity and stability.[2] When used with an appropriate filter, usually available from the manufacturer of the photovoltaic cell, the spectral response approximates that of the CIE relative luminous efficiency function. For values of luminance greater than about 1 fl, the photocurrent may be read directly with a microammeter [1]–[3], [6], [7].

2.10.1.2.2 Photoemissive type: this type is required if the luminance range to be measured includes values much less than 1 fl, but instruments suitable for use at high luminance levels are also available.[3] The relatively wide range of variation of spectral sensitivity among phototubes makes luminosity correction by fil-

[1] The most commonly used visual photometer is the Macbeth Illuminometer. See Section 2.10.7.

[2] An example of a widely used photometer of the photovoltaic type is the Footlambert Meter, Weston Model 759. See Section 2.10.7.

[3] An example of a photoemissive type of photometer is the Spectra Brightness Spot Meter. See Section 2.10.7.

ters a problem that requires a specific solution for individual phototubes [2], [3], [6].

2.10.1.2.3 Thermopile type: This type requires a very sensitive low-impedance detector and very precise ambient temperature control. The response time of this detector is relatively long. Because of the nonselective nature of the detector, the spectral transmittance of the luminosity filter alone determines the accuracy with which the *CIE* relative luminous efficiency function is duplicated [5].

2.10.1.2.4 Photoconductive type: This type has not been applied to the measurement of screen luminance because of surface instability.

2.10.2 Calibration of Photometers:

2.10.2.1 Calibration of visual photometers: Illuminate perpendicularly a uniform surface of known 45°0° reflectance[4] R with light from a standard lamp[5] at a distance D, operated at the prescribed voltage to provide the rated intensity I in the specified direction. The luminance B in a direction 45° from the perpendicular is computed as follows:

$$B = \frac{RI}{D^2}.$$

When I is expressed in candles, D in feet, and R as a decimal fraction, B will be given in footlamberts. A sufficient number of distances must be chosen to provide luminance calibration over the desired range. Position the photometer so as to view the center of the surface at an angle of 45° and at a sufficiently close distance so that the field of view is entirely filled by the illuminated surface. The instrument reading must not vary with change of distance to the surface, and this condition must be verified experimentally.

Brightness matches are more readily made when the colors to be matched are similar; therefore, for improved accuracy the reference standard and the working standard lamp in the photometer must be altered by filters to a color approximately that of the fluoresence of the phosphor to be measured. The color-correcting filter should never be used to alter the chromaticity of the source being measured. Suitable filters can be calibrated by the National Bureau of Standards or other standards laboratories.

Complete luminance standards including standard lamp and filter combinations can be obtained from the National Bureau of Standards, Washington 25, D. C.

[4] The 45°0° luminous directional reflectance (for brevity called reflectance) is the ratio of the luminous flux from a specimen illuminated at an angle of 45° and viewed perpendicularly by the *CIE* standard observer to the luminous flux from the standard magnesium oxide layer, similarly illuminated and viewed. Refer to the ASTM Method E97, "Method of Test for 45°0° Directional Reflectance of Opaque Specimens."

[5] Standard lamps calibrated in intensity and in color temperature may be purchased from the National Bureau of Standards, or an uncalibrated lamp may be calibrated for these characteristics by the National Bureau of Standards.

2.10.2.2 Calibration of physical photometers: Physical photometers calibrated with steady light are subject to inaccuracies when used to measure pulsating light sources. Physical photometers when used to measure cathode-ray tubes should therefore be calibrated by use of cathode-ray tubes operated at specified raster conditions and measured by means of a visual photometer.

Since the cathode-ray tube, when displaying a raster, is a pulsating light source, and since the wave shape of the pulses is a function of many factors including size and shape of the raster, and the fraction of the raster which the photometer views, phosphor persistence characteristic, etc., the calibration of the physical photometer from a cathode-ray tube as described above is completely valid only for use of the photometer under identical conditions of tube operation with the same tube type, same fraction of the raster viewed, and with any other physical factors kept unchanged if the change of these factors affects the photometer readings.

If a visual photometer is not available for calibration of the physical photometer, the physical photometer may be calibrated by means of a cathode-ray tube screen luminance standard consisting of a standard lamp and filter combination obtainable from the National Bureau of Standards, Washington 25, D. C.

Physical photometers calibrated by means of steady light sources may indicate erroneous values of luminance for cathode-ray tube screens displaying rasters because the photometer and the human eye do not integrate the pulsating light of a cathode-ray tube in the same manner. Therefore, when using a physical photometer calibrated by means of a steady light source to measure the luminance of a cathode-ray tube screen displaying a raster, care should be taken to insure that the pulsating light does not result in erroneous readings. For example, it has been found that extreme variations in raster size, or extreme variations in the per cent of the raster which the instrument views may affect the readings of the instrument. If changes in such conditions are observed to produce unusual changes in instrument readings, the manufacturer of the instrument should be consulted as to the suitability of the instrument for measurements under the particular desired conditions.

The readings of physical photometers on pulsating light sources such as cathode-ray tubes may not be independent of the distance between the photometer and the light source. This has been found to be the case for certain physical photometers, where varying the distance between the photo receptor and the cathode-ray tube significantly changes the wave shape of the light pulses reaching the receptor.

2.10.3 Tube Operation: The operating conditions are adjusted in accordance with the published data for the particular cathode-ray tube under test. Particular attention must be given to the conditions listed below:

1) The tube must be operated with a sharply focused beam that is scanning a raster at the specified line and frame frequencies.

2) Raster linearity must be carefully adjusted throughout the screen area.
3) Vertical retrace blanking must be employed.
Note: Frequently it is desired to measure power input to the cathode-ray tube screen at the same time that screen luminance is being measured, in order to compute screen efficiency. If the method of measuring screen luminance described here is used, in which vertical retrace blanking is employed, it should be be remembered that the current to the screen is not continuous but is actually pulsating because of the presence of the vertical retrace blanking. Suitable metering should be employed to measure the average value of this pulsating current.
4) The raster must be carefully adjusted to the designated size.
5) For a color picture tube, the desired chromaticity must be selected and maintained during luminance evaluation. (See Section 2.11.)

2.10.4 Measurement of Screen Luminance: Using a calibrated visual photometer, the tube is adjusted to produce the desired value of screen luminance. An average of several readings taken by an experienced observer should be used for each luminance level.

When numerous readings on tubes of the same type are required, a physical photometer may be used to gain a higher degree of repeatability with a minimum of operator variation. After a tube is measured by the visual photometer, it may be used as a secondary standard for physical photometers. The reading obtained from a physical photometer at a given fixed distance from the tube will calibrate the photometer for other tubes of the same type at the same luminance level if the operating conditions of the secondary standard are the same as those used during its measurement. (Refer to Section 2.10.2.2.)

It should be stressed that calibrations of physical photometers by a secondary cathode-ray tube standard may be valid only on tubes of the same type and when all of the conditions listed in Section 2.10.3 are duplicated.

2.10.5 Stray Emission: With the tube operated in accordance with its published data, with the control grid biased beyond cutoff and with the scan adjusted to stated test conditions, the luminance of the screen due to stray emission is measured. Since the luminance caused by stray emission will, in general, be at a very low level, a very small amount of ambient illumination will affect the measurement. (See Section 1.4).

2.10.6 Cathode Illumination: With the specified heater potential applied and with all other electrode voltages zero, the luminance of the screen due to light from the cathode is measured. Since the luminance caused by light from the cathode assembly will, in general, be at a very low level, very small amounts of ambient illumination will affect the measurement. (See Section 1.4.)

2.10.7 Some Commercial Photometers:

Macbeth Illuminometer, Leeds and Northrup Company, Philadelphia, Pa.

Luckiesh-Taylor Brightness Meter, General Electric Company, Schenectady, N. Y.

Footlambert Meter, Weston Model 759, Weston Instruments, Division of Daystron, Inc., Newark, N. J.

Spectra Brightness Spot Meter, Photo Research Corporation, Hollywood, Calif.

2.11 Chromaticity of Screen Luminescence

The radiation gamut of cathode-ray tube phosphor screens extends from the ultra-violet through the visible region into the infrared. The radiation characteristic can be accurately given in the form of a spectral-energy-distribution (SED) curve, which is a plot of screen relative radiant emittance per unit wavelength as a function of wavelength. The color of the radiation from cathode-ray tube screens which have radiation in the visible region can be expressed by a set of chromaticity coordinates, preferably the x and y coordinates of the CIE system [10], [11], [13], [22].

The CIE coordinates can be transformed into those of a number of other coordinate systems of specialized importance, among which are:

1) Correlated color temperature and minimum (at least) perceptible color difference units from the black body locus [11], [13].
2) Dominant wavelength and purity [10], [11].
3) Uniform chromaticity scales [11], [12].

2.11.1 Introduction: The fundamental method of describing the character of the screen emittance of a cathode-ray tube is by means of a spectral-energy-distribution (SED) curve, from which chromaticity can be computed. Two cathode-ray tube screens may have radiant outputs appearing to be of the same color even though their SED curves are somewhat different. This anomaly is the result of the manner in which the eye of an observer weighs the contributions of energy from different parts of the spectrum.

Because the spectral sensitivities of the eyes of actual observers differ, an average of normal observers was obtained and designated the *standard observer.* The spectral-response functions for the standard observer were adopted by the Commission Internationale de l'Eclairage (CIE) in 1931. The tristimulus functions for an equal energy source are designated $\bar{x}$, $\bar{y}$, $\bar{z}$. These functions relate the amounts of three theoretical primary colors required by the standard observer to form a stimulus equivalent to a spectral color at each wavelength. The color of a light source is described by specifying the amounts of these primaries required to match the color; these amounts are designated by the tristimulus values X, Y, and Z. The chromaticity of the color is usually described in terms of the chromaticity coordinates x, y, and z, obtained by dividing X, Y, and Z by

the sum $X+Y+Z$. Since the three functions thus obtained always sum to unity, only two of them need to be specified, usually x and y [10], [11], [13].

The chromaticity of the luminescence of a cathode-ray tube screen is determined most accurately by a spectroradiometer. A photoelectric tristimulus colorimeter is another instrument that is often used. The spectroradiometer yields the SED curve of the screen, and the data are treated as indicated above to yield chromaticity coordinates. The photoelectric tristimulus colorimeter, when properly calibrated, yields a set of values that can be converted directly into chromaticity coordinates. The calibration of the colorimeter must be made, for best results, with a cathode-ray tube screen whose tristimulus values have been computed from spectroradiometric data.

The validity of the measurement of chromaticity of self-radiant sources is often subject to question. The methods described are believed to be the most accurate currently available, but methods of better accuracy continue to be sought. Reference [44] contains much valuable information on the general subject of color measurement, and [45] contains excellent material on the difficulties encountered in making color measurements of self-luminous sources.

2.11.2 Spectroradiometers:

2.11.2.1 Instrumentation: The instrumentation consists of a dispersion device, a detector of radiant energy, and a readout device. The dispersion device is a spectrometer of either the prism or the grating type. The detector is a photosensitive device, such as a multiplier phototube, having adequate stability and sensitivity throughout the visible spectrum. The readout device must have high sensitivity, be suitable for use with the detector employed, and accurately measure the detector output. In the design of the detector-readout combination consideration must be given to the fact that the input light may be pulsating. In the measurement of screens that have appreciably different persistence over different portions of their spectral output the lack of correspondence between the detector and the human eye in the integration of pulsating light may introduce errors. The magnitude of these errors is believed to be small. For detailed descriptions of instrumentation, refer to the literature on spectroradiometers [26]–[33].

2.11.2.2 Calibration:

1) Wavelength calibration—The wavelength scale is established by means of spectral emission lines of gas-discharge lamps containing helium, or cadmium and mercury.

2) Photometric linearity—The linearity of the photometric scale is checked over the range to be used by applying the inverse-square law to a small incandescent lamp operated at constant voltage at measured distances from a diffusing glass placed over the entrance slit of the spectrometer.

Precaution: Care must be taken to adequately eliminate reflections of light from various objects, such as the equipment itself, the operator's clothing, etc.

3) Spectral-radiance calibration—The spectral radiance is calibrated by using an incandescent lamp operated at a known color temperature to obtain the equivalent of a Planckian radiator having a relative spectral-radiance distribution $U_{\lambda s}$. CIE source A, 2854°K is usually satisfactory.[6]

The lamp is placed near the entrance slit of the spectrometer, a heat shield being placed between the lamp and the spectrometer, if necessary. The light from the lamp is allowed to fall on a diffusing material, such as white vitrolite glass,[6] having a known spectral-reflectance distribution $\rho\lambda$. A blue filter of known spectral transmittance $\tau\lambda$, such as a Corning daylight filter,[6] is placed over the entrance slit of the spectrometer. The distance between the lamp and diffuser is adjusted so as to obtain instrument readings of the same order of magnitude as will be obtained from the cathode-ray tube. The spectrometer slits are made as narrow as consistent with reasonable detector sensitivity. Instrument readings $d_{\lambda s}$ are obtained for equal intervals not greater than 10 millimicrons over the wavelength range 400 to 700 millimicrons. The calibration factors $C\lambda$ at each wavelength are then computed from the equation

$$C\lambda = \frac{U_{\lambda s}\rho\lambda\tau\lambda}{d_{\lambda s}}.$$

2.11.2.3 Measurement:

1) Tube operation—The operating conditions are adjusted in accordance with the published data for the particular cathode-ray type under test. The tube is placed so that the screen is parallel to the plane of the entrance slit, and centered with respect to it. The tube must be sufficiently close to the spectrometer so that the optical system is completely filled with light.

2) Spectrometer operation—The Corning daylight filter used in calibration must be removed from in front of the entrance slit. The spectrometer must be shielded from ambient light and from screen light reflected from surfaces external to the tube. (See Section 1.4.) The same slit widths are used as those used during calibration of the instrument. Spectrometer readings of spectral energy $d_{\lambda t}$ must be obtained at sufficiently close wavelength intervals to characterize the spectral-radiance distribution of the source being measured.

2.11.2.4 Computation:

1) Spectral-energy distribution—The relative spectral-radiance distribution $U_{\lambda t}$ of the test source is computed from the relation

$$U_{\lambda t} = d_{\lambda t}C_{\lambda}.$$

[6] Calibrated lamps and vitrolite glass tiles are available from the National Bureau of Standards, Washington 25, D. C. Filters can be calibrated by the National Bureau of Standards or other standards laboratories.

2) Chromaticity—The relative tristimulus values are computed as follows:

$$X = \sum_{\lambda_2}^{\lambda_1} U_{\lambda t}\bar{x}$$

$$Y = \sum_{\lambda_2}^{\lambda_1} U_{\lambda t}\bar{y}$$

$$Z = \sum_{\lambda_2}^{\lambda_1} U_{\lambda t}\bar{z}$$

$$x = \frac{X}{X + Y + Z}$$

$$y = \frac{Y}{X + Y + Z}$$

$\lambda_1 = 700$ millimicrons

$\lambda_2 = 400$ millimicrons.

2.11.3 Photoelectric Tristimulus Colorimeter:

2.11.3.1 Instrumentation: The tristimulus colorimeter consists of one or more photodetectors, filters to modify their spectral response to the desired tristimulus functions, and either a readout device or analog computer. The tristimulus functions most generally used for cathode-ray tube measurements are the CIE functions $\bar{x}$, $\bar{y}$, and $\bar{z}$. The $\bar{y}$ and $\bar{z}$ functions are approximated each by a filter-photodetector combination. The $\bar{x}$ function is approximated by either one or two filter-photodetector combinations. Some instruments have simple readout devices that provide data from which the x and y coordinates are computed. Other types of colorimeters incorporate analog circuits, the outputs of which provide chromaticity coordinates directly. For detailed descriptions of instrumentation, refer to the literature [18], [19], [24], and [25]. It must not be assumed that published filter and photodetector spectral data are accurate enough to design a precise colorimeter. For most accurate results, the spectral transmittance of filters and the spectral response of photodetectors must be measured and tailored to each other to obtain accurate duplication of the CIE tristimulus functions. In the measurement of screens that have appreciably different persistence over different portions of their spectral output, the lack of correspondence between the detector and the human eye in the integration of pulsating light may introduce errors. The magnitude of these errors is believed to be small.

2.11.3.2 Calibration: The recommended calibration standard is a cathode-ray tube whose color has been determined by means of a spectroradiometer. Three-filter nonanalog colorimeters require one standard tube, and its color should be similar to the color to be measured. Four-filter nonanalog colorimeters require two standard tubes that are sufficiently different in chromaticity to bracket the color to be measured.[7]

To calibrate the analog-computer type of instrument, the amount of light from the standard tube impinging on the detectors is adjusted or the electrical gain is changed until the output x and y values of the colorimeter agree with the x and y values of the calibrated tube.

Colorimeters without analog circuits require the determination of a series of constants K_n. Readings are made of the photodetector-filter combinations on calibrated tubes having known x, y, and z coordinates.

The constants are determined from the following equations:

1) Three-filter colorimeter [19]. This requires at least one calibrated tube.

$$\text{a)}\quad K_1 = \frac{G(x - 0.167z)}{yA}$$

$$\text{b)}\quad K_3 = \frac{Gz}{yB}$$

$$\text{c)}\quad K_2 = 0.167K_3,$$

where

G = output through green ($\bar{y}$) filter,
A = output through amber ($\bar{x}$) filter,
B = output through blue ($\bar{z}$) filter.

2) Four-filter colorimeter [18]. This requires at least two calibrated tubes whose screens have emittances of different spectral-energy distribution.

$$\text{a)}\quad \epsilon_1 = \frac{G_1}{y_1} \qquad \epsilon_2 = \frac{G_2}{y_2}$$

$$\text{b)}\quad K_1 = \frac{x_1\epsilon_1 D_2 - x_2\epsilon_2 D_1}{A_1D_2 - A_2D_1}$$

$$\text{c)}\quad K_3 = \frac{G_1z_1}{y_1B_1} = \frac{G_2z_2}{y_2B_2}$$

$$\text{d)}\quad K_4 = \frac{x_2\epsilon_2 A_1 - x_1\epsilon_1 A_2}{A_1D_1 - A_2D_2},$$

where

D = output through blue (x) filter.

2.11.3.3 Measurement: The operating conditions are adjusted in accordance with the published data for the particular cathode-ray-tube type under test. The viewed raster size and luminance are adjusted to be the same as that used for calibration. Raster linearity and focus are carefully adjusted. The readings on the analog computer type of instrument will be the CIE co-

[7] Cathode-ray tubes can be calibrated for chromaticity by the Electronic Industries Association Laboratory, 32 Green St., Newark, N. J.

ordinates as discussed above. The nonanalog instruments yield a set of readings that are entered into the following equations to give x and y:

1) Three-filter instruments:

$$x = \frac{K_1A + K_2B}{K_1A + G + (K_2 + K_3)B}$$

$$y = \frac{G}{K_1A + G + (K_2 + K_3)B}.$$

2) Four-filter instruments:

$$x = \frac{K_1A + K_4D}{K_1A + K_4D + G + K_3B}$$

$$y = \frac{G}{K_1A + K_4D + G + K_3B}.$$

2.12 Screen-Persistence Characteristic

2.12.1 Introduction: The persistence or decay characteristic of a cathode-ray-tube screen is usually described by a curve relating the light output of the screen to the time after excitation ceases. The persistence may also be specified as the time required for the phosphorescent light to decay to an arbitrary percentage of the value of the light at the instant of cessation of excitation. This time can range from a fraction of a microsecond to several seconds.

The recommended procedure for measurement of persistence is the following:

> The tube whose screen persistence is to be measured is operated with a pulsed, focused, undeflected spot. A multiplier phototube with suitable sensitivity and spectral response picks up the light from the tube screen. An oscilloscope or electromechanical recorder is used to display the output of the multiplier phototube as a function of time.

2.12.2 Instrumentation:

2.12.2.1 Pulse generator: The pulse duration and pulse repetition rate required from the pulse generator depend upon the persistence characteristic of the phosphor to be measured. The fall time of the pulse must be short compared to the decay time being measured. The pulse must have a flat top of duration greater than the build-up time of the phosphor being measured. The combination of pulse repetition rate and pulse duration must be such that essentially complete decay of the phosphor occurs between pulses. With extremely short-persistence phosphors, it may be necessary to supply pulses as short as 0.01 μsec in order to obtain a sufficiently high pulse repetition rate to get a reasonable display of the persistence characteristic on the oscilloscope. With long-persistence phosphors it may be possible to use pulse durations of the order of 100 msec. If many phosphors with different persistence characteristics are to be measured, it is convenient to have a pulse generator that will cover the entire range of pulse durations and repetiton rates. In order to cover the normal range of cathode-ray-tube cutoffs, it is convenient to have the pulse amplitude variable from zero to about 100 volts peak-to-peak. For certain phosphors with long build-up times and persistence characteristics, the pulse generator is not required; instead, the cathode-ray-tube beam may be turned on and off by switching the bias on the cathode-ray-tube control electrode from one value to another either manually or by suitable automatic means.

2.12.2.2 Multiplier phototube: A multiplier phototube load having satisfactory stability and sensitivity must be used.[8] The power supply must provide variable control of the multiplier gain. The multiplier-phototube load resistor should have a value low enough so that the RC constant of the circuit is short compared to the decay time being measured, yet large enough to provide the desired sensitivity. To determine that the value of the resistor is sufficiently low, it should be reduced until further reduction results in no change to the measured decay characteristic. To cover the range of persistence characteristics from very short to very long persistence phosphors, it is desirable to be able to adjust the load resistor over the range from 70 ohms to 10 megohms. It may be desirable to isolate the radiant output of individual components of multicomponent screens by means of suitable optical filters.

2.12.2.3 Display device: An oscilloscope with vertical deflection amplifiers of suitable bandwidth and gain and a horizontal sweep of appropriate rate for the phosphor under test is used. For very short persistences a bandwidth of at least 10 Mc is required, and for long persistences sweep times of the order of two minutes or more for full scan are desirable. For long-persistence phosphors a suitable electromechanical recorder may be substituted.

2.12.3 Method of Measurement: The operating conditions are adjusted according to the published data for the particular cathode-ray tube under test to produce an undeflected focused spot on the screen. A voltage waveform having the following characteristics is applied to the control grid:

a) Duration must be such as to allow complete luminance build-up of the phosphor during the on period of the tube.

b) Amplitude must be such as to give a suitable deflection on the oscilloscope without screen or photomultiplier saturation.

c) Repetition frequency must be low enough to permit the screen under test to decay to a negligible luminance level between pulses.

The peak pulsed-anode current must be adjusted to the same value as the steady-state anode current at

[8] Type 6217 has been found satisfactory for most purposes.

which it is desired to make the measurement. The peak anode current may be conveniently measured by an oscilloscope.

The oscilloscope, operated at suitable sweep speed and sensitivity, is connected across the load resistor of the multiplier phototube, and the decay curve is read directly or recorded by means of a camera.

2.12.4 Precautions:

1) Since the persistence characteristic of most phosphors is more or less dependent on excitation conditions, the spot-size variation from tube type to tube type or spot-size variation due to different conditions of focus on the same tube may affect the measured value of persistence.

2) Care must be exercised in choosing the multiplier-phototube load resistor to avoid capacitive loading at the high frequencies, which increases the apparent time constant of the decay curve.

3) Filters for isolating the radiant output of components of multicomponent screens must be chosen carefully to eliminate, so far as possible, all output from components other than the one to be tested.

4) In some applications the cathode-ray tube is operated under such conditions that the output of the phosphor may neither build up completely nor decay completely between successive excitations of the same point on the phosphor screen. It may be desired to make measurements of the variations of screen radiant output under such conditions of excitation, in which case the pulse width and repetition rate should be chosen to duplicate as closely as possible the successive excitations resulting from the scanning in the intended application.

5) Care must be taken that other circuit parameters, including the input capacitance of the cathode-ray tube under test, do not alter the pulse waveform supplied by the pulse generator to such an extent that the pulse waveform appearing at the control grid under test no longer meets the criteria specified for the pulse generator in Section 2.12.2.1.

6) The calibration and linearity of both the horizontal and vertical ordinates of the display device must be ascertained.

2.13 Large-Area Contrast

2.13.1 Introduction: Subjective evaluations of cathode-ray-tube displays, including not only television pictures, but also radar and oscilloscope displays, show that the utility of the display and the amount of information which can be obtained from it are directly related to the degree of contrast that can be obtained from the cathode-ray tube. Low contrast ratios reduce the number of individual luminance levels discernible in the display, and so reduce the amount of information obtainable from the display. Two general classifications of contrast can be described, large-area contrast, and detail contrast.

The large-area contrast of a television picture tube is the ratio of the values of luminance at two widely separated areas on the screen with the screen excited by the electron beam to a particular luminance level in one of the two areas, and with the beam biased off in the other area. Detail contrast is the ratio of the values of luminance of adjacent picture elements, and is intimately associated with resolution.

The large-area contrast of a picture tube is affected by many factors, and widely differing values may be obtained for any given tube, depending upon the conditions of measurement. The large-area contrast of any given tube is not necessarily uniquely determined by the physical structure of the tube or the general electrical conditions under which it is operating at the time of measurement. Factors such as raster size and shape, picture content, reflected light from objects external to the tube, chromaticity and luminance of ambient illumination influence large-area contrast measurement. The large-area contrast of a tube is also affected by the amount of light from the excited portions of the screen that reaches the unexcited areas, and by undesired excitation of the dark areas of the picture by stray electrons.

Picture content, such as relative shape, sizes and location of excited and unexcited areas, greatly affects large-area contrast. For example, in Fig. 1(a), where the screen is excited directly only at the center, the large-area contrast obtained will be considerably higher than in Fig. 1(b), where the screen is excited everywhere except at the center.

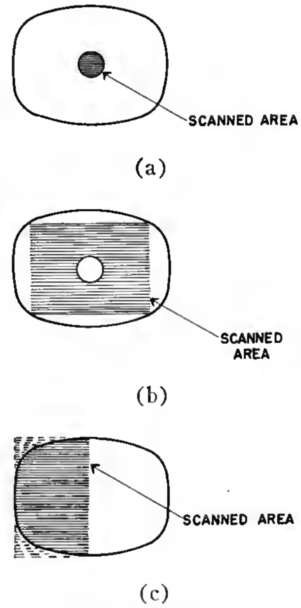

Fig. 1—Rasters used for various types of large area contrast measurements.

2.13.2 Method of Measurement: A recommended method of measuring large-area contrast *for comparison purposes* follows.

The cathode-ray tube is operated in accord with its published data and with a scanned raster at the speci-

fied line and frame frequencies, and of normal height and only half normal width for the tube type under test. The raster is positioned with either the right or left edge passing through the screen center. [See Fig. 1(c).] The ambient illumination must be kept low enough so that any further reduction of the ambient illumination will not change the measurement.

The large-area contrast is given by the formula

$$\text{Large Area Contrast} = \frac{L_1 - LO_1}{L_2 - LO_2},$$

where

L_1 = luminance at Position 1 near the center of the excited part of the screen,
LO_1 = luminance at Position 1 with beam biased off,
L_2 = luminance at Position 2 near the center of the unexcited part of the screen,
LO_2 = luminance at Position 2 with beam biased off.

The values of luminance L_1 and L_2 are measured with the grid bias adjusted for the desired raster luminance.

2.13.3 Precautions:

1) The measurement will be greatly affected by the amount of overscanning. It is important in comparing tubes that the raster be adjusted, as in Fig. 1(c), to just touch the top, bottom and one side of the viewing area of the tube with only the corners overscanned. If measurements are made under other degrees of overscanning, the amount of overscanning must be specified.

2) The foregoing standard method of measurement is specified to be made with low ambient illumination because of the extreme difficulty of specifying and controlling all characteristics of the ambient light. However, it is recognized that picture tubes are normally used under conditions where the ambient illumination does affect the large-area contrast. It should be realized that measurements made under low ambient illumination may not correlate at all with measurements made under higher ambient illumination.

2.14 Resolution

2.14.1 Introduction: The resolution of a cathode-ray tube may be stated in several different ways. Frequently only the limiting resolution is stated, although it is only one aspect of the over-all resolution capability of a tube. The most complete description of the resolution capability of a cathode-ray tube can be stated in terms of the sinewave response of the tube. Several other methods of determining some of the resolution capabilities of a cathode-ray tube are sometimes used. Some of these yield somewhat more information than just the limiting resolution value, but they do not yield as complete and readily interpreted data on the over-all resolution as is obtainable using the sinewave response method. A description of some of these other methods of measuring resolution may be found in Part 10.

2.14.1.1 Limiting resolution: The limiting resolution of a cathode-ray tube is the maximum number of black and white lines distinguishable either vertically or horizontally. It can be measured by means of relatively simple test patterns, such as RETMA Resolution Chart 1956. It should be recognized that the limiting resolution has, in general, no fixed relation to the subjective sharpness of television images.

2.14.1.2 Sinewave response: The subjective judgment of the quality of picture detail depends on the ability of the moving spot to reproduce an electrical input step transition as a step transition in light intensity. It can be seen that the total length of the light-intensity transition is equal to the diameter of the beam spot, while the shape and gradient of the transition depend on the energy distribution in the beam spot. The light-transition curve that is generated when the beam spot responds to an electrical step transition is therefore a sensitive measure of over-all resolution. However, the shape of a *single* light transition is not a sufficient specification of image quality since the shape of the transition is affected when several transitions occur within a distance less than the maximum spot diameter. Thus it can be seen that a good measure of the over-all resolution or ability of a cathode-ray tube to reproduce video information consists of some measure of the ability of the cathode-ray tube to reproduce electrical step transitions of various frequencies as light transitions.

The light-transition curve that results from a series of electrical step-transition inputs is, however, difficult to analyze. If instead of a step-transition input a sinewave input is used, the amplitude of the output light variation as a function of the input sinewave frequency may be analyzed. The curve of output amplitude vs input sinewave frequency is a description of the over-all resolution capability of the tube. In addition, this curve may be analyzed to yield a single figure of merit describing the resolution capability of the tube. This figure of merit N_e has gained some acceptance as a satisfactory single value which can be used in some instances in place of a descriptive curve.

2.14.2 Measurement of Limiting Resolution by Means of Resolution Patterns: The cathode-ray tube to be tested is operated in accordance with the published data for the tube type under test. A video signal such as that produced by the RETMA Resolution Chart 1956, consisting of a pattern of lines, wedges, or other test designs, is fed to the control grid or cathode of the tube under test. Particular attention must be paid to setting video pattern size, raster interlace, linearity, anode voltage and beam current or luminance at the specified values during the test. Care must also be taken that the control grid or cathode bias and the video signal level are adjusted to the desired light output or beam current, so that the beam is just cut off in the black portions of the picture, and so that the gray scale included in the chart is correctly reproduced. The focusing electrode or magnet is then adjusted to produce the desired focus.

The pattern is visually examined, and the point in the wedge or test design where the black and white lines merge and become indistinguishable is noted. The number of lines limiting resolution at this point is read from the test chart. The "lines" of limiting resolution (the number of just distinguishable black and white lines) must not be confused with the scanning "lines." Measuring technique employing the RETMA Resolution Chart 1956 is discussed in detail in the "IRE Standards on Television, 1950 (50 IRE 23.S1)."

Normally in using resolution patterns the raster size is adjusted to just fill the useful screen area. If the limiting resolution as read on the test chart appears to be greater than the maximum number of lines represented on the chart, the raster size may be reduced in its horizontal or vertical dimension or both. If this is done, the limiting horizontal resolution number read on the chart is inversely proportional to the reduction in raster size in the horizontal dimension, and the limiting vertical resolution is inversely proportional to the reduction in raster size in the vertical dimension.

2.14.2.1 Precautions and limitations in the measurement of limiting resolution by means of resolution patterns: In any system of cascaded devices each device has a limiting effect on the resolution. In the television system the camera tube, the pass band of the radio-frequency and video-frequency amplifiers, and the display device or cathode-ray tube each place a restriction on the attainable resolution of the system. In addition, the nature of the scanning system employed causes a fundamental difference in the horizontal and vertical structure of the image.

2.14.2.1.1 Vertical-resolution limitations: Basically, the vertical resolution (limiting resolution in the vertical direction, *i.e.*, perpendicular to the scanning lines) cannot exceed the number of active scanning lines in the raster. If the picture elements are displaced so that each element falls directly between two adjacent scanning lines, none of the picture elements is distinguishable. Experience has shown that, for random patterns, the vertical resolution is limited to about 70 per cent of the active scan lines. Thus, in the American television system, only $490 \times 0.70 = 343$ television lines can be resolved in the vertical direction on the television image, regardless of the quality of the spot obtained in the cathode-ray tube involved.

2.14.2.1.2 Horizontal-resolution limits: The horizontal resolution (limiting resolution in the horizontal direction, *i.e.*, parallel to the scanning lines) is limited by the maximum number of changes in electron-gun drive voltage that can occur during each line scan. The rapidity with which the voltage variation can be produced depends, in turn, on the bandwidth of the video channel employed.

2.14.2.1.3 System limitations: In making a resolution measurement by means of a test chart the raster size should be increased in the horizontal direction to insure that the video system is not limiting the resolution. If the number of distinguishable black and white lines in the resolution wedge does not increase, the resolution is being limited by the video system instead of by the cathode-ray tube under test.

If the system is limiting the resolution, the technique of reducing the raster size may be employed, as discussed in Section 2.14.2. If this is done, a point on the resolution wedge well below that which represented the video system limitation is selected. The raster size is then reduced until the lines of the wedge just become indistinguishable at this selected point. The tube resolution is calculated by multiplying this line number by the raster size-reduction factor.

2.14.3 Measurement by Means of Sine-Wave Response: As discussed in Section 2.14.1, there are fundamental limitations in describing resolution by means of limiting resolution only. The sine-wave-response technique avoids these limitations by giving a more complete description of the beam spot in terms of its ability to reproduce on the screen of the tube sinusoidal input signals of various frequencies and amplitudes. These input signals are displayed on the screen of the tube as a pattern of lines having a sinusoidal variation of light intensity, perpendicular to the lines, as in Fig. 2. The equipment, technique, and computation used in making sine-wave-response measurements on a cathode-ray tube is sufficiently complicated to make a description in general terms of very limited value. The entire method is treated in detail in Schade [43], and familiarity with this reference is essential in an understanding of this method.

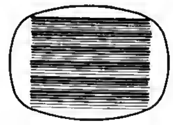

Fig. 2—Sine wave variations in light intensity during measurement of line number.

2.14.3.1 Instrumentation: The measurement of the sine-wave response requires three basic pieces of test apparatus: 1) a suitable signal generator to supply sine-wave modulation of the desired frequencies (approximately 10 to 30,000 cps is satisfactory for most uses) to the control electrode of the tube under test; 2) a suitable beam-deflection and energy-supply system capable of producing a noninterlaced raster of about 500 lines at the desired operating conditions on the tube under test; 3) a microphotometer consisting of a photomultiplier tube mounted in an optical system incorporating a narrow slit. This is used to measure and record amplitudes of the light intensity of the sinusoidal pattern on the cathode-ray tube screen. The details of an apparatus to measure sine-wave response is given in the literature [43].

2.14.3.2 Method of measurement: The tube is oper-

ated in accordance with the published data. A video sine-wave input is applied to the control element of the tube. The amplitude of the signal must be kept small (about 15 per cent) compared with the steady background signal to prevent waveform distortion by the nonlinear transfer characteristic of the cathode-ray tube and to reduce variations in the cathode-ray-tube beam diameter, which is, in general, a function of beam current.

A noninterlaced raster of about 500 lines is scanned on the tube under test. The frequency of the sinusoidal modulating signal applied to the control element of the tube under test is adjusted so as to be an integral multiple of the field frequency plus or minus a few cps (approximately 10).

$$f_m = pf_f \pm f_d$$

f_m = modulating frequency

p = an integer

f_f = field frequency

f_d = drift frequency.

It can be seen that this will result in a pattern of lines having a sinusoidal variation of light intensity, and that these variations in light intensity will appear to drift across the scanning lines at a frequency f_d.

The slit of the microphotometer is aligned with the pattern of lines having a sinusoidal variation of light intensity and thus allows the photomultiplier tube to observe a small portion of the sine-wave pattern as it drifts past the slit at a frequency f_d. The amplitude of the sinusoidal output of the photomultiplier therefore corresponds to the amplitude of variation in luminance of the sine-wave pattern on the screen. Signals having values of f_m corresponding to several values of p (typically from about $p=0$ to $p=50$) at a constant value of f_f and f_d are applied to the control element of the tube under test, and the corresponding amplitudes of photomultiplier output are recorded for each value of f_m.

The sinusoidal variation of light intensity that the photomultiplier tubes sees as the sine-wave pattern drifts past the slit is not continuous. It is actually a modulation envelope on a pulse carrier wave having a frequency n_r, the total raster-line number. Elimination of unwanted sidebands ("beats") requires that both the modulating frequency f_m and the optical pass band of the cathode-ray tube under test be limited to one half the carrier frequency n_r. The optical pass band f_c equals one half the line number N_c, and $N_c \leqq n_r$. Thus f_m, the modulating frequency, must be limited to one half the horizontal deflection frequency, and the vertical raster dimension V (the dimension perpendicular to the scanned lines of the raster) must be adjusted so that the cathode-ray tube under test will not resolve the line number n_r. This condition is obtained by reducing V to one half the value giving a "flat" field with zero modulation.

Since the spot of the cathode-ray tube may not be symmetrical, measurements should be made with scanning applied first in one direction, and then in at least one other direction, usually perpendicular to the direction first chosen.

2.14.3.3 Computation: The value of f_m is directly proportional to the line number N, and N can be computed by the formula

$$N = 2(n_r/n_s)(f_m/f_f)/V,$$

where

N = line number, lines per inch,
n_r = total raster line number,
n_s = unblanked or visible raster lines,
V = vertical raster dimension.

A complete explanation will be found in the literature [43].

The outputs of the photomultiplier at the various values of f_m are all normalized to the output at the lowest value of f_m, and these values of relative output are plotted against the corresponding values of line number N. This curve of relative photomultiplier output vs line number is a complete description of the resolution capabilities of the cathode-ray tube under test.

If it is desired to express the resolution as the single value N_e, this number is determined as follows.

The outputs of the photomultiplier at the various values of f_m are all normalized to the output at the lowest value of f_m, and these values are then squared. By plotting these squared values of output vs line number and integrating the area under the curve, the figure of merit N_e may be established.

N_e is the line number corresponding to the boundary of a rectangle superimposed on the plot having unity height and the same area as that under the original curve, as shown in Fig. 3.

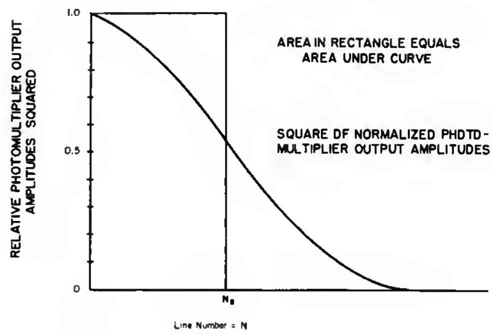

Fig. 3—Method of determining figure of merit N_e.

Tests made with pictorial subjects, including test patterns, have shown that the subjective relative sharpness and definition of cathode-ray-tube images with equal N_e ratings are substantially equivalent even though the limiting resolution may differ as much as 3 to 1.

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Part 3

Gas Tubes

Subcommittee 7.3

Gas Tubes

H. H. Wittenberg, *Chairman* (1961–1962)
A. W. Coolidge, *Past Chairman* (1956–1960)
M. A. Townsend, *Past Chairman* (1955)

J. H. Burnett	D. E. Marshall	L. W. Roberts
E. J. Handley	W. Minowitz	W. W. Watrous
R. A. Herring	G. R. Riska	A. D. White

1. Introduction

1.1 General Precautions

The characteristics of gas tubes differ radically from those of (high) vacuum tubes. Because the tube voltage drop is nearly independent of the magnitude of current conducted and the grid in general has no control after the discharge has started, the current through the tube is determined primarily by the applied voltage and the circuit parameters.

The precautions of Part 1, Introduction, are applicable, including the method of establishing equivalent datum points for ac and dc filament supplies.

Mercury vapor is often used in gas tubes. The pressure of the vapor in the tube affects the electrical characteristics and is a rapidly varying function of the condensed-mercury temperature.

In testing mercury-vapor tubes care must be taken to control condensed-mercury temperature in order to obtain reproducible results. In general, with increase of temperature the tube voltage drop will decrease, the peak inverse voltage that the tube will withstand will decrease, the critical grid voltage will become more negative, and the recovery time will become longer.

It should be noted that the time required to attain the desired condensed-mercury temperature is usually longer than the cathode heating time.

In testing the maximum peak inverse voltage or the control characteristics precautions must be taken to insure that no condensed mercury exists in the upper parts of the tube. This is generally accomplished by a preheating process in which only cathode heating power is applied to the tube. The tube must be maintained in an upright position, away from air drafts, in order to prevent mercury from recondensing in the upper part of the envelope.

2. Hot-Cathode Gas-Tube Tests

2.1 Filament or Heater Electrical Characteristic[1]

[1] For test methods, see Part 1.

2.1.1 Precaution: For tubes with high filament current, it may be necessary to correct for voltage drop in the socket when the voltage is measured at the socket terminals.

2.2 Control-Characteristic Tests

The critical grid voltage of a gas tube is a function of anode voltage, and is usually presented in the form of a curve with instantaneous anode voltage as ordinate. Data for plotting the curve are obtained as follows.

2.2.1 Critical-Grid-Voltage Test: With voltage applied to the filament or heater, and with sufficient direct voltage applied to the control grid to prevent conduction, direct voltage is applied to the anode through a resistance that will limit the current during conduction to a suitable value. The grid series resistance is usually zero, but should in no case be large enough to affect the test result. The control-grid voltage is gradually made more positive (or less negative) until breakdown occurs, at which time the critical grid voltage is observed. Where the critical grid voltage is positive, appreciable grid current may flow before breakdown. The grid voltage must then be read at the grid terminal.

The condensed-mercury temperature of mercury-vapor tubes should be held to the desired value which should be recorded with the results. This value is dependent upon operating conditions and tube structure, but is ordinarily between 20° and 50°C.

For tubes having shield grids, the shield-grid potential is an additional parameter, and should be held at a specified value, usually zero.

2.2.1.1 Precaution: It may be necessary to use a small capacitor between each grid and the cathode to prevent possible anode-voltage surges from affecting the grid voltage.

2.2.2 Critical-Anode-Voltage Test: With voltage applied to the filament or heater, the grid or grids, if any, are held at desired voltages and the anode voltage increased until an irreversible change in anode current occurs. The anode voltage at this point is the critical anode voltage. Sufficient resistance must be inserted

into the anode circuit to limit the current during conduction to a suitable value.

2.3 Emission Tests

In a tube of given design tube voltage drop is primarily a function of emission. Lower tube voltage drop is, therefore, an indication of higher emission. The electron-emission quality of a gas tube is commonly tested by measurement of the tube voltage drop at an anode current sufficiently high to indicate cathode capability. This is sometimes done under conditions of continuous-current conduction. An intermittent-conduction method is more accurate because of the reduction of cathode heating or cooling by the arc current. Formation of a cathode spot, as indicated by an intermittent reduction in the arc drop, indicates faulty cathode operation.

In these tests any grids having separate terminals should be connected to the anode either directly or through a current-limiting resistor. The value of resistance should be specified, since the tube voltage drop is affected by the current distribution between the grid and the anode.

Since the tube voltage drop in mercury-vapor tubes varies with vapor pressure, the permissible range of condensed-mercury temperature should be controlled within narrow limits.

2.3.1 Direct- and Alternating-Voltage Methods: With the desired voltage applied to the filament or heater, with an anode supply voltage sufficient to cause firing, and with the cathode current limited to the desired value by a series resistor, the voltage drop is measured by a voltmeter if direct voltage is used, and by an oscillograph or other suitable means if alternating voltage is used. In the latter case the voltage is read at the instant the current is at its peak value.

2.3.2 Intermittent-Voltage Method: With the desired voltage applied to the filament or heater, with the cathode current limited to a desired value by the circuit, and with the anode voltage to be applied intermittently at a duty cycle such that appreciable heating or cooling of the cathode does not occur, the tube voltage is measured by an oscillograph or other suitable means at the instant when the current has peak value. The period of conduction should be of sufficient duration to insure that the tube is sufficiently ionized to make the measured tube voltage drop independent of the conduction period.

2.4 Grid-Current Tests

The critical grid current of a thyratron is a function of the anode voltage and includes positive-ion, electron, and leakage currents. The interelectrode-capacitance charging currents may also be of importance. These grid currents are usually of the order of microamperes. They are generally measured by reading the voltage drop across a resistor in series with the grid.

2.4.1 Critical-Grid-Voltage Method (for Tubes Having Negative-Control Characteristics): With the desired value of alternating anode voltage, the grid-supply voltage is made sufficiently positive to maintain the desired anode current. With a resistance R_g in series with the grid, as shown in the circuit of Fig. 1, the grid supply voltage is then made more negative until conduction ceases. This voltage is denoted by E_{cc}. The measurement is then repeated with a different resistance R_g' which should be sufficiently low so that there is negligible voltage drop across it due to preconduction grid currents and sufficiently large to avoid varitions in grid-supply voltage resulting from grid current during conduction. The second measured value is denoted by E_{cc}'. Since the critical grid voltage can be assumed to have the same values in both measurements, the grid current flowing just before conduction is

$$i_{cc} = \frac{E_{cc} - E_{cc}'}{R_g - R_g'}.$$

Normally R_g' is much less than R_g and can be neglected. The frequency should be low enough so that the effect of tube and circuit capacitances can be neglected.

The primary electron emission from the grid is a function of the grid temperature. In order that the full value of total grid current may be measured, the grid must be heated by operating the tube at the desired values of anode current for an adequate time immediately before the measurements are made.

In this test the shield grid is connected to the cathode. Other grid-current tests may be made with direct voltage on the shield grid.

2.4.1.1 Precaution: This method may be inaccurate if the positive-ion current to the grid resulting from the preconduction current in the thyratron is large compared to the electron-emission current from the grid and if the resistance of R_g is too great.[2] In this case use a self-protecting electronic microammeter.

2.5 Fault-Current Test (Surge-Current Test)

The ability of a tube to withstand a forward fault current without excessive damage is tested in a circuit in which the tube is connected in series with a high-current tube such as an ignitron, as shown in Fig. 2. By control of the ignitron, the desired peak value of half-wave current can be passed through the tube for a definite time.

2.6 Operation Test

Gas tubes may be given an operation test by operating them in conventional rectifier circuits. Fig. 3 shows a typical full-wave single-phase circuit; Fig. 4 shows a typical three-phase zig-zig circuit as connected for thyratrons. In the testing of diodes the grid circuits are eliminated.

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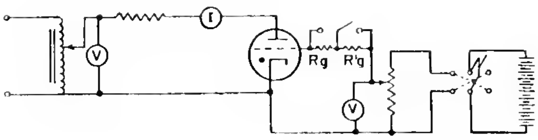

Fig. 1—Circuit arrangement for measuring critical grid current.

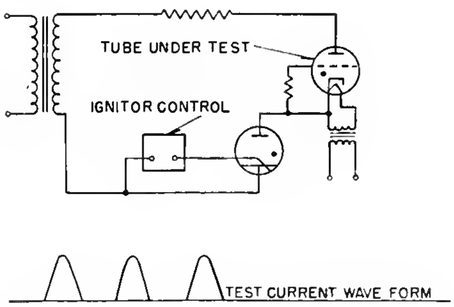

Fig. 2—Circuit arrangement for fault-current test.

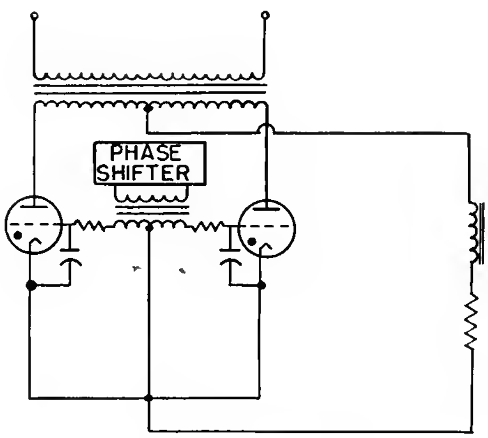

Fig. 3—Circuit arrangement for single-phase-rectifier operation test.

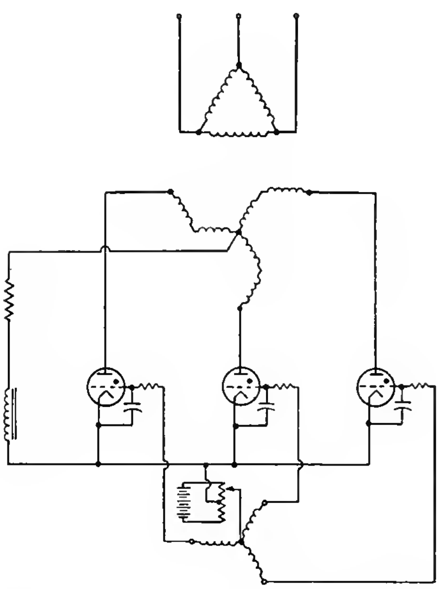

Fig. 4—Circuit arrangement for three-phase-rectifier operation test.

The tube is operated at a definite temperature for a definite time to observe the frequency of arc-back and whether grid control is lost.

The severity of the test depends upon a number of tube factors such as those illustrated in Fig. 5. These and other important factors are:

1) Peak anode current.
2) Peak inverse anode voltage.
3) Current at the beginning of commutation.
4) Inverse voltage immediately after commutation.
5) Commutating time.
6) Arc-back current.

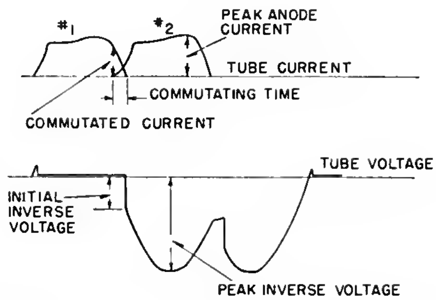

Fig. 5—Typical three-phase-rectifier waveforms.

The magnitude of these factors depends upon the following circuit parameters:

a) Number of phases of the rectifier.
b) Type of filtering circuit.
c) Transformer impedance.
d) Type of load circuit. (The power may be absorbed either by resistor or by a reverse-EMF load. Examples of the latter are a battery, a capacitor, or a motor, which can augment the arc-back current.)
e) Firing time in grid-controlled or ignitor-controlled tubes.

The correlation of test results may be difficult unless the tests are made under conditions that are identical with respect to the foregoing parameters.

For convenience, this test is often made with a full-wave single-phase rectifier operating into a resistance load, such as a water resistor. The influence of factors 3), 4), and 5) is then minimized. In order to make the test comparable in severity with a polyphase test, it is usually necessary to increase the test voltage.

2.6.1 "Cheater-Circuit" Test: It is possible to construct test circuits, called "cheater circuits," that will simulate the current and voltage waveforms of the circuit described in the last paragraph of Section 2.6. These circuits usually consist of high-current low-voltage sources of power to supply the forward current through the tube under test, and high-voltage low-current sources to apply inverse voltage after the end of the conduction period. Synchronous switching methods are used to switch from one source to another at the proper time in the cycle. The object of such circuits is to re-

duce the energy requirements and the equipment cost. This usually requires that the high-voltage source of power be of relatively small capacity and high impedance. The current through the tube on occurrence of arc-back is therefore relatively small. Under such conditions, it is possible that the rate of occurrence of arc-back may be reduced over that which would be obtained from a power source of lower impedance. By the use of sensitive arc-back indicators, the small reverse currents that flow when the tube arcs back can be indicated. Such a method of testing requires careful correlation of test results with those obtained under actual service conditions.

Fig. 6 shows one form of "cheater circuit." The switching is done by a rotating commutator or by a suitable thyratron circuit.

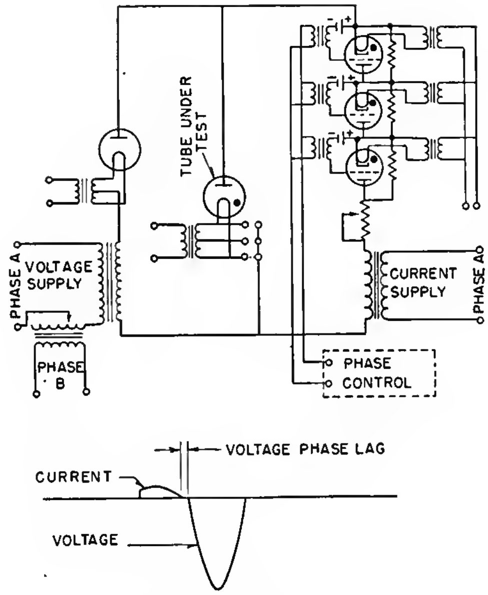

Fig. 6—Typical "cheater-circuit" rectifier test circuit.

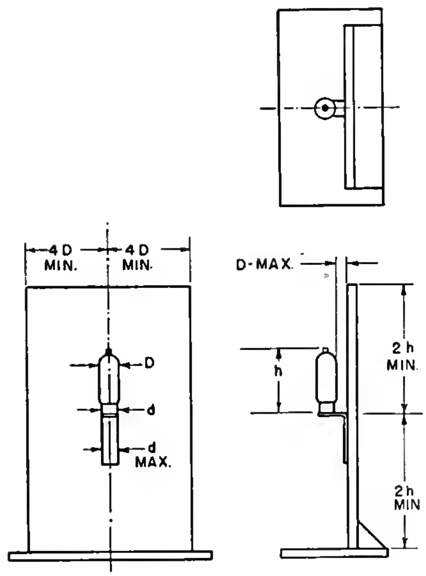

Fig. 7—Test mounting for mercury-vapor-tube temperature tests.

2.7 Thermal Tests for Hot-Cathode Mercury Tubes

The operating temperature of a hot-cathode mercury-vapor tube, cooled by air convection, depends on the power generated in the tube and the efficiency of the air cooling. The latter depends on the ambient air temperature and on the air flow past the cooling surfaces of the tube.

The time required for the temperature of a tube to rise from a low ambient value to its minimum desired operating value depends on the filament power and on the ambient temperature. Tube arc losses are not involved because the tube should not be allowed to conduct before it is heated to the minimum operating temperature.

In order to fix the conditions under which cooling efficiency is measured, it is necessary to establish the mounting method. A suggested arrangement is shown in Fig. 7. Baffles or other means should be used to prevent extraneous drafts.

2.7.1 Method of Measuring Condensed-Mercury Temperature: The preferred method of measuring the condensed-mercury temperature is by means of a thermocouple in close contact with the tube in the region in which the mercury is condensing. The location of the thermocouple is usually immediately above the base of a glass tube or on the radiator of a metal tube. The thermocouple wire must be of small enough diameter to make the error caused by heat conduction negligible.

2.7.2 Test for Rate of Condensed-Mercury Temperature Rise: The condensed-mercury temperature of the tube is measured as a function of time, starting at the time the heater circuit is closed. The heater voltage should be held constant, preferably at the lowest desired rated operating value. The data obtained may be plotted as a curve of temperature rise above ambient temperature vs time.

2.7.3 Test for Equilibrium Condensed-Mercury Temperature Rise: With the filament or heater voltage at the highest operating value and with maximum value of average anode current, the condensed-mercury temperature rise at equilibrium is measured for various ambient temperatures. The data obtained may be presented as a curve of condensed-mercury temperature vs ambient temperature.

2.7.4 Cathode-Heating-Time Test: The cathode heating time may be determined with the aid of an optical pyrometer or by the direct measurement of the tube drop.

2.7.4.1 Optical-pyrometer method: The desired filament or heater voltage is applied to an initially cooled tube, and the temperature of the emitting surface is monitored with an optical pyrometer. The time between the application of the filament or heater voltage and the time when the cathode temperature reaches operating temperature is the cathode heating time.

2.7.4.1.1 Precaution: When it is difficult or impractical to obtain the cathode heating time in accordance with 2.7.4.1, the cathode heating time may be ob-

tained in accordance with 2.7.4.2. Conditions leading to this situation are:

a) Emitting surface not visible.
b) Operating temperature too low for pyrometer reading.
c) Emitting surface exposed to direct radiation from a heater at much higher temperature.
d) Cathode heating time very short.

2.7.4.2 Tube-drop method: With the filament or heater voltage adjusted so that the equilibrium cathode temperature is equal to an acceptable operating temperature, the tube voltage drop at rated peak current is measured in accordance with 2.3.2. This tube voltage drop, designated "A", is utilized in determining the cathode heating time.

After the tube has been allowed to cool, the desired filament or heater voltage is applied. At some later time, the tube drop is measured at rated peak current in accordance with 2.3.2. The minimum value of elapsed time required to produce a tube drop equal to tube drop "A" described in the paragraph above is the cathode heating time.

2.7.4.2.1 Precaution: To avoid damage to the cathode when determining cathode heating time in accordance with 2.7.4.2, it is recommended that in the first trials the elapsed time be made great enough to assure that the cathode has reached operating temperature. Measurements are made at successively shorter times in order to arrive at the cathode heating time. The tube should be allowed to cool after each exploratory measurement.

2.8 Recovery-Time Test

The time from end of conduction until grid control is reestablished is called the recovery time. This time is a function of a number of variables, among the more important of which are voltage and impedance of the grid circuit, anode current previous to extinction, anode voltage, and (in mercury-vapor tubes) the temperature of the envelope.

2.8.1 Inverter Method: The time necessary for recovery may be measured by connecting two of the tubes in an inverter circuit, as shown in Fig. 8. A common practice is to use a tube of known characteristic in this circuit, together with the tube to be tested.

The test circuit is similar to a full-wave single-phase rectifier, except that the load, a dc generator in this case, can cause current to flow through the tubes. Triggering pulses are applied to the grids of the tubes so that each tube fires a short time before the ac voltage on its anode becomes negative. The dc generator in combination with the smoothing reactor forces current through the tubes in opposition to the transformer voltage. Thus, during the greater part of the ac cycle, power flows from the generator to the ac source.

The operation of the circuit may be visualized by referring to the circuit of Fig. 8 and to the voltage and

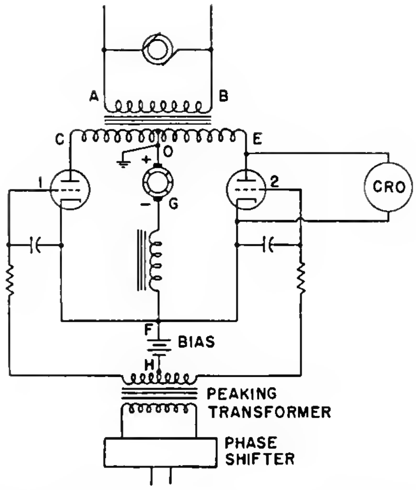

Fig. 8—Inverter circuit for measuring recovery time.

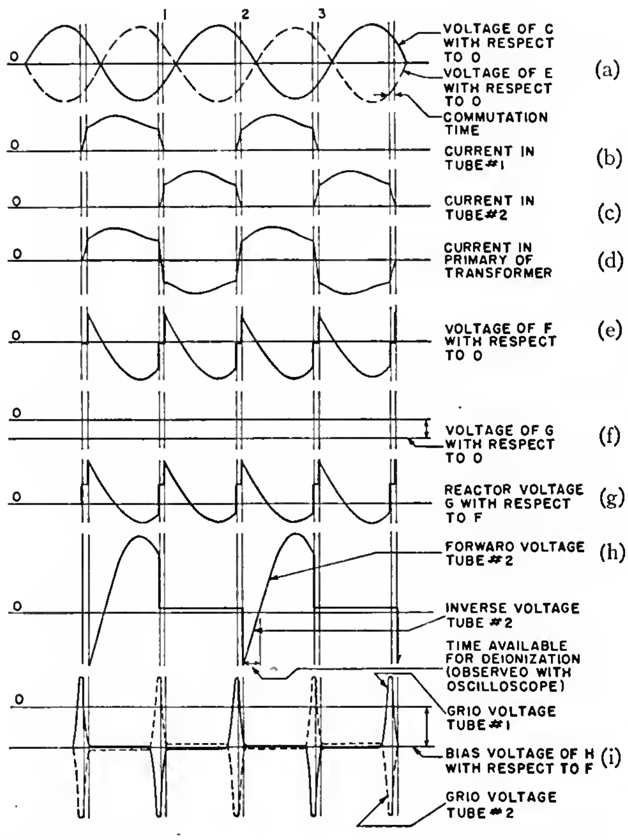

Fig. 9—Typical waveforms appearing in the inverter circuit for measuring recovery time.

current waveforms of Fig. 9. Assume current to be flowing in tube No. 2, as shown in Fig. 9(c). Since the voltage drop in the tube is small relative to the applied voltages, the voltage of point F in Fig. 8 may be assumed to be equal to that at point E as long as tube No. 2 is conducting. From Figs. 9(a) and 8 it can be seen that, with tube No. 2 conducting, the anode of tube No. 1 is positive with respect to its cathode during the latter portion of the conduction cycle of tube No. 2, and tube No. 1 is kept from conducting only because its grid bias is more

negative than the critical value. At point 2 of Fig. 9(a), the grid of tube No. 1 is pulsed as shown in Fig. 9(i). Tube No. 1 immediately starts to conduct. However, because of the leakage reactance in the transformer, current through the transformer secondary DE cannot stop instantly, and similarly the current through tube No. 1 and transfomer winding CD cannot build up instantly. Therefore, for a finite period known as the commutation time, both tubes are conducting. Since point C of Fig. 8 is now positive and point E is negative, tube No. 1 continues to conduct and tube No. 2 is extinguished. The voltage from F to D across the inductance and the generator is as shown in Fig. 9(e). It is the sum of the rectified alternating voltage and the direct voltage.

The voltage of the anode with respect to the cathode of tube No. 2, shown in Fig. 9(h), is then the voltage of E with respect to D Minus the voltage of F with respect to D.

The method of adjustment of this circuit is as follows:

1) Set the dc bias at the value specified for the test.
2) With the dc generator disconnected from the inverter, set the dc generator voltage to a value low with respect to its operating value.
3) Adjust the alternating voltage until the specified forward peak voltage is obtained.
4) Adjust the phase shifter to fire the tubes somewhat in advance of the desired phase.
5) Connect generators to the circuit and raise the alternating voltage until the tubes start to conduct an appreciable current.
6) Adjust the dc-generator voltage and phase of firing together to obtain the desired average tube current. (Phasing later in the cycle or raising the anode current will change the inverse voltage and require adjustment of the generator voltage.)
7) Run for the specified time and note failures to commutate.
8) Repeat the test for longer or shorter times of negative anode voltage until a duration of negative anode voltage is established for the specified number of recovery failures per unit time.

The inverse and forward voltages on the tubes depend on the alternating voltage. The average amount of current flowing is a function of the direct voltage and of the angle of lag of the grid voltage at a given alternating-anode voltage.

When a tube fails to recover within the available time, the direct current rises to a high value. A circuit breaker must therefore be provided in a dc circuit to protect the tubes. A choke should be used to smooth the current and to decrease the rate of rise of direct current on recovery failure, and thus to prevent the direct current from increasing sufficiently to damage the tubes within the operating time of the circuit breaker.

Commutation failure may result from failure of arc initiation in one of the tubes. The initiating grid voltage should therefore be made sufficiently positive for an adequate time to insure reliable initiation of the arc.

2.8.2 Capacitor-Discharge Method: If a charged capacitor is connected to apply a negative voltage to the anode of a gas tube that is conducting, the discharge in the tube will cease. The current previously flowing through the tube now charges the capacitor in the opposite direction, causing the anode voltage to change from negative to positive at a rate dependent on the parameters of the charging circuit. If the ion density decreases sufficiently while the anode is negative, the grid can prevent conduction when the tube voltage becomes positive.

Fig. 10 shows a circuit using this principle to measure recovery time. The tube under test is conducting direct current of the desired value. The capacitor, which is connected in parallel with the load resistor, charges in such polarity that the terminal connected to the tube anode is negative. When the switch is closed the anode voltage becomes negative, and the discharge therefore ceases. The capacitor then discharges through the load resistor. The waveform of the voltage across the tube is shown in this figure. With this method, the recovery time is arbitrarily considered to be the time taken by the anode voltage to rise to a value equal to the tube voltage drop, with the capacitance adjusted to the smallest value at which the tube stops conducting a desired percentage of a number of trials. The recovery time can be calculated from the formula

$$t = RC \ln \frac{2E_{bb} - E_a}{E_{bb} - E_a},$$

where E_a is the tube voltage drop, and E_{bb} is the supply voltage.

(*Note:* When the tube voltage drop is less than 10 per cent of the anode-supply voltage, it may without serious error be assumed to be zero. The formula for recovery time then simplifies to $t = 0.7RC$.)

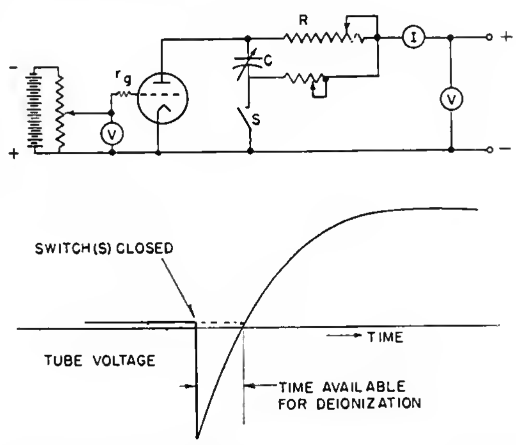

Fig. 10—Circuit arrangement for measuring recovery time by capacitor-discharge method.

2.8.3 De-ionization-Time Test: The time from end of conduction until the tube regains all preconduction characteristics is called de-ionization time. Recovery times vs all variables are measured and displayed as families of curves. The time at which the final curve becomes asymptotic to the precondition value is the de-ionization time.

2.9 Thyratron Ionization-Time Test

With the filament or heater voltage at the desired value and the tube temperature adjusted to a value within the desired range, a voltage, either direct or alternating, is applied to the anode through a resistance. With the grid biased to a voltage substantially more negative (or less positive) than the critical grid voltage, an essentially rectangular pulse of voltage of variable duration is applied in the grid circuit. This pulse should be of such magnitude that the resultant grid voltage is less negative (or more positive) than the critical grid voltage by a specified amount and, when alternating anode voltage is used, should occur at the peak of the anode-voltage wave. Conduction should take place within the time of the pulse duration, as indicated by a cathode-ray oscillograph. The ionization time is the time from the application of the grid pulse to the time at which the tube voltage drop falls to a specified value.

2.9.1 Precaution: The time of rise of the applied pulse should be short compared with the ionization time to be be measured.

3. Cold-Cathode Gas-Tube Tests

3.1 General Precautions

The characteristics of cold-cathode tubes are affected by several factors that are not, in general, common to other types of tubes. The more important factors are discussed in the following sections.

3.1.1 Illumination: The breakdown voltage and ionization time of tubes not having an opaque coating may be markedly affected by the wavelength and intensity of illumination. As illumination on the tube is increased, the breakdown voltage and ionization time decreases. For consistent test results, it is advisable that all characteristics be determined in a range of illumination where the tube characteristics are not materially affected by the illumination.

Cold-cathode tubes, in general, have a memory effect with regard to illumination. With a sudden decrease in illumination intensity, the effect of the previous higher value often tends to persist, but decreases with time.

3.1.2 Storage and Handling: When a tube is allowed to stand idle, ionization time and breakdown voltage may increase. This effect is especially apparent when the tube is subsequently operated at low levels of illumination. After the tube has conducted for a few seconds, the ionization time and breakdown voltage revert to approximately their original values. The maximum changes due to storage ordinarily take place in from twenty-four hours to one week.

3.1.3 Ambient Temperature: In tubes containing only inert gases the effect of temperature changes on the tube characteristics is usually negligible, except for the tube voltage drop of voltage reference and regulator tubes.

3.1.4 Rapid Test Methods: Cold-cathode gas tubes may have comparatively long ionization and recovery times. In measuring such characteristics as breakdown voltage and transfer current it is therefore necessary that the tube be allowed to come to equilibrum under the test conditions. In automatic or rapid test methods it is recommended that test voltages be applied for a period of not less than 0.1 second.

3.1.5 Capacitance Effects: In the measurement of transfer current major errors may be introduced if capacitance effects are ignored. Circuit capacitance between the starter and the cathode can supply an appreciable transfer current not indicated on the current meter. This current is caused by the discharge of this capacitance through the tube, the voltage falling by an amount equal to the difference between the breakdown voltage and voltage drop of the starter. It is important, therefore, that this effect be eliminated in this test by placing the starter resistor adjacent to the starter.

3.1.6 Preconditioning Current: In order to eliminate the effect of storage on starter breakdown voltage in testing certain narrow-range tubes, a preconditioning current is sometimes required. It is recommended that this preconditioning of the electrode to be tested take place not more than 30 seconds prior to making the test, that the value of the preconditioning current be at least 10 per cent of rated value of the tube, and that the preconditioning period be not less than 0.1 second and not more than 5 seconds.

3.2 Breakdown-Voltage Tests

3.2.1 Anode Breakdown-Voltage Test: All electrodes, except the anodes, should be held substantially at cathode potential before breakdown. The anode circuit should contain a resistor having sufficient resistance to limit the current, after breakdown to a suitable value. A positive anode voltage is applied and increased until the tube conducts current. The minimum voltage required to cause breakdown is measured.

3.3.2 Starter Breakdown-Voltage Test: This test is similar to that in Section 3.2.1, except that the anode and the starter are interchanged.

3.3 Anode-Voltage-Drop Tests

In three-electrode tubes ordinarily both anode voltage drop and starter voltage drop are measured. When anode voltage drop is being measured, the starter may be allowed to float or may be connected to the cathode through sufficient resistance to limit the starter current to not more than 10 per cent of the anode current.

In measuring starter voltage drop the anode should be approximately at cathode potential. The current to the anode should not exceed 10 per cent of the starter current.

3.3.1 Direct-Current Method: A positive direct potential is applied to the anode (or starter) through sufficient resistance to limit the current to the desired value. The anode-to-cathode (or starter-to-cathode) voltage drop is then measured.

3.3.2 Alternating-Current Method: An alternating voltage is applied to the anode (or starter) through sufficient resistance to limit the peak current to the desired test value. The anode-to-cathode (or starter-to-cathode) voltage drop is then measured at peak current with a cathode-ray oscillograph or by other suitable means.

3.4 Transfer-Current Test

A positive potential is applied to the anode through sufficient resistance to limit the current after breakdown. A positive voltage is applied to the starter through sufficient resistance to limit the starter current to less than the transfer value. The starter current is then gradually increased until conduction takes place to the anode. The starter current just prior to breakdown is measured.

The starter resistor must be adjacent to the starter (see Section 3.1.5).

3.5 Voltage-Regulator-Tube Regulation Test

The difference between the maximum and the minimum anode voltage drop is determined as the anode current is varied over the desired range.

3.6 Drift Rate (of a Voltage Reference or Regulator Tube)

A positive direct potential is applied to the anode through the proper resistance to obtain a specified current. With the tube in continuous operation at this current, measurements of tube voltage drop are made at specified intervals of time. The data are plotted, and a smoothed curve is drawn by averaging short time variations. Drift rate at any time is the slope of the curve at that time. The ambient temperature should be maintained constant throughout the test.

3.7 Repeatability (of a Voltage Regulator or Reference Tube)

A positive direct potential is applied to the anode through the proper resistance to obtain a specified current, and the tube voltage drop is measured after a specified time interval. The anode voltage is removed for a specified time and reapplied successively for a desired number of applications. The tube voltage drop is measured at the specified time after each application. The lack of repeatability is the maximum difference between the measured values of the tube voltage drop.

3.8 Temperature Coefficient of Voltage Drop (of a Glow Discharge Tube)

A positive direct potential is applied to the anode through the proper resistance to obtain a specified current. With the tube in continuous operation at this current, the tube voltage drop is measured at two specified ambient (or envelope) temperatures. The temperature coefficient is the quotient of the change of tube voltage drop (excluding any voltage jump) by the change of ambient (or envelope) temperature.

Note: In measuring the temperature coefficient sufficient time must be allowed for tube temperature to reach equilibrium before each measurement. It should be indicated whether the coefficient is taken with respect to ambient or envelope temperatures.

3.9 Voltage Jump (Arising from Changes in Current)

Current through the tube under test is changed at approximately a specified rate between two specified values. The tube voltage drop is monitored on an instrument having specified band-pass characteristics. The magnitude of any voltage jump is the magnitude of any abrupt change or discontinuity in tube voltage drop.

In performing this test, if the current cycle is repeated periodically, the observer or the monitoring instrument must ignore voltage jump arising from the abrupt change in current at the beginning and end of each cycle.

Part 4

Microwave-Duplexer Tubes

Subcommittee 7.3.1

Microwave-Duplexer Tubes

L. W. Roberts, Chairman (1957–1962)
K. Garoff, *Past Chairman* (1955–1956)

I. Birnbaum
N. Cooper
L. Gould
H. Heins
W. J. Kearns
F. Klawsnik
A. Marchetti
F. McCarthy
I. Reingold
J. Schussele
R. Scudder
E. Vardon
R. Walker
S. Miller

1. Introduction

Duplexing devices are used primarily in pulsed radar systems to enable a single antenna to perform both the transmitting and the receiving function.

Four basic conditions must be satisfied if the device is to operate satisfactorily. First, the duplexer must connect the transmitter to the antenna, and disconnect it from the receiver during the period of transmission. Second, the duplexer must thoroughly isolate the receiver from the transmitter during the transmission period to prevent damage to the sensitive elements of the receiver by power leakage past the duplexer. Third, the duplexer must rapidly disconnect the transmitter and connect the receiver to the antenna after the transmission period. Finally, the duplexer should absorb as little power as possible during operation.

The duplexer usually consists of a gas-discharge device in a resonant microwave structure (although ferrites are finding increasing application) and the associated microwave circuitry. The gas-discharge device forms a class of microwave-duplexer tubes known individually as ATR (anti-transmit receive), pre-TR, TR and dual TR tubes.

Duplexing circuits can be divided into two classes: branched and balanced circuits. The usual branched circuit consists of two Tee junctions in the transmission line, with the ATR tube mounted on the Tee nearest the transmitter, and the TR tube (and pre-TR tube, if used) mounted on a Tee placed an odd number of quarter wavelengths from the ATR tube. A common type of balanced circuit consists of a dual TR tube mounted between two 3-db hybrid directional couplers.

Duplexer tubes are subjected to the high incident power of the transmitter and extremely low power reflected from distant targets. Therefore, evaluation of the tube performance must include the determination of its properties in a quiescent state (low-power characteristics), in an ionized state, and during the transition period between the two states (high-power characteristics).

Low-power-level test procedures are used to measure the properties of the duplexer tubes that determine the characteristics of the nonionized tube. These measurements are used to determine equivalent RF circuit parameters and to indicate directly or indirectly the nonionized characteristics of the tubes. Furthermore, these circuit parameters directly affect many of the characteristics of the tube in the ionized state.

High-power-level measurements are used to examine the charateristics of the duplexer tube while the RF gas discharge is being initiated, sustained, and extinguished. The results are not expressed in terms descriptive of the discharge itself, but in parameters associated with the operating functions of the tube.

The test procedures described in the following sections are used for determining and predicting the performance of the microwave duplexer and the various types of gas-discharge tubes that are integral elements of the duplexer.

Reference is made in the test procedure to various matched elements. These elements are matched to the characteristic impedance of the particular transmission line used in making a specific measurement (that is, the VSWR is low, 1.05 or less).

In some cases, more than one method of measurement is indicated in describing a particular tube characteristic. The standard or preferred method in all such cases is marked with an asterisk. The alternate method will provide, however, test results which are sufficiently accurate for most applications and may be considerably simpler to perform.

2. Low-Level Radio-Frequency Measurements

2.1 *Tuning Susceptance (ATR Tubes)*

The measurement of susceptance indicates how well the tube is tuned at a specified frequency; a perfectly tuned tube would have zero susceptance. Although a direct measurement of resonance frequency would be desirable, the low Q of an ATR tube makes this ap-

proach inaccurate. Two methods of measuring tuning susceptance follow.

Method A:[1] The tube, in a series mount, is inserted into a transmission line between a matched generator and a matched power detector and power meter, as in Fig. 1. With the generator set at the specified frequency, the power P_i incident upon the tube and the power P_0 transmitted to the matched power meter, are determined. The normalized susceptance b of the tube can then be computed from

$$b^2 = \frac{K(1+2g)^2 - 4g^2}{4(1-K)},$$

where

g = normalized conductance of the tube (see 2.2 for measurement technique),
$K = P_0/P_i$,
P_i = the power incident upon the tube,
P_0 = the power transmitted to the matched power meter.

To determine K, a dummy tube (*i.e.*, one in which the window is covered by a metallic short circuit) is placed into the mount, and the output level at the matched terminating detector is set by adjustment of the variable attenuator. The tube under test is then inserted into the mount, and the calibrated attenuator is readjusted until the detector output is equal to that obtained by the dummy tube. Instead of employing a dummy tube, the entire ATR tube mount may be removed from the test line during the initial setting of the variable attenuator. The difference in attenuator readings gives the power ratio K. Because this is a power-substitution method it is essential that the monitored signal-generator power level remain constant throughout the measurement.

The power reflected by the tube is generally comparable in magnitude to the incident power. A fixed attenuator or isolator (see Fig. 1) is placed in the line to isolate the large tube mismatch from the signal generator and its monitoring circuits.

Method B: The tube, in a series mount, is inserted into the transmission line, as shown in Fig. 2. The susceptance is measured by comparing the phase of the standing wave before the tube with that of a tube that is known to be resonant at the specified frequency (standard tube). The normalized susceptance is then

$$b = \frac{(1+2g)}{2} \tan \frac{4\pi\Delta l}{\lambda g},$$

where

b = normalized susceptance of the tube,
g = normalized conductance of the tube (see 2.2 for measurement technique),
Δl = shift in position of the standing-wave minimum when the standard tube is replaced by the sample tube,
λg = guide wavelength (expressed in same units as Δl).

This can be expressed approximately as

$$b = (1+2g)\frac{2\pi\Delta l}{\lambda g}$$

for small values of Δl.

The procedure for measuring Δl is to determine the position of the voltage-standing-wave minimum, first with the standard tube in the mount, and then with the tube under test.

An alternate measurement of Δl can be made with a magic Tee or other suitable type of RF impedance bridge. The tube, in the series mount, is placed into the line as shown in Fig. 3.

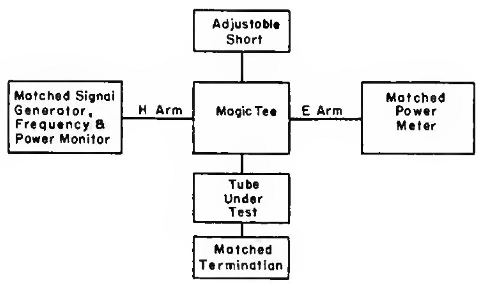

Fig. 3—Block diagram for susceptance measurement by an impedance bridge.

The standard tube is inserted into the mount, and the adjustable short is varied until the matched detector output is a minimum. The tube under test is then placed into the mount and the short is readjusted for minimum power output from the detector. The change in the position of the adjustable short for the observed minima between the standard tube and the tube under test is Δl.

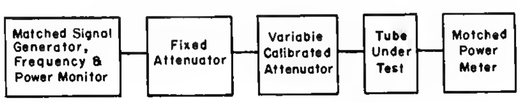

Fig. 1—Block diagram for susceptance measurement by Method A.

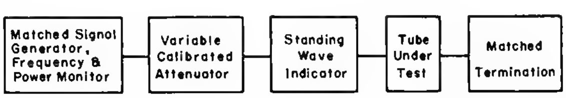

Fig. 2—Block diagram for susceptance measurement by Method B.

2.2 *Normalized Equivalent Conductance (ATR Tubes)*

The magnitude of the normalized equivalent conductance of a tube is a factor in determining the duplexer loss for a given power-source impedance.

[1] Standard method.

Method A:[1] The tube, in a series mount, is inserted into a transmission line, as shown in Fig. 2. The maximum value of the voltage-standing-wave-ratio S_r occurs at the resonance frequency of the tube, at which the normalized susceptance b is zero. Therefore,

$$g = \frac{1}{S_r - 1} \qquad (S_r > 1).$$

As a rule, however, g is measured at the specified frequency employed in the equivalent susceptance measurement. If the susceptance of the tube is small at the specified frequency ($b \ll g$), the normalized equivalent conductance is

$$g \approx \frac{1}{S_s - 1} \qquad (S_s > 1),$$

where S_s = VSWR at the specified frequency.

As in 2.1 (Method A), the use of a calibrated attenuator is recommended.

Method B: The tube in a series mount is inserted between a matched generator and a matched termination, as shown in Fig. 1. With the generator set at the resonance frequency (that frequency at which the susceptance is zero), the power P_i incident upon the tube and the power P_0 transmitted to the matched termination are determined. The resonance frequency is obtained by varying the frequency of the signal generator until the power at the termination is a minimum. The power incident upon the tube should be maintained constant in searching for the resonance frequency. The normalized equivalent conductance is

$$g = \frac{1}{2(\sqrt{m} - 1)},$$

where

$$m = \frac{P_i}{P_0}.$$

As in 2.1 (Method A), the use of a calibrated attenuator is recommended. This method is subject to appreciable error if the signal generator is not tuned closely to the resonance frequency of the tube.

2.3 Loaded Q (ATR Tubes)

Duplexing loss is a function of ATR susceptance. Since the tube is used over a frequency band, the measure of loaded Q, Q_L, gives an indication of the loss at the band edges.

Method A:[1] Q_L is defined in terms of the rate of change of the normalized susceptance b with frequency f and is expressed as

$$Q_L = \frac{f_r}{2(1 + g)} \frac{\Delta b}{\Delta f},$$

where

f_r = resonance frequency,
g = normalized conductance of tube (see 2.2 for measurement technique),
$\Delta b/\Delta f$ = slope of the linear portion of the b vs f curve around f_r.

The procedure for performing this measurement consists of determining b at some frequency f above or below f_r in order to obtain the slope $\Delta b/\Delta f$. The linear relationship between b and f exists within ± 1 per cent of f_r. Two methods for measuring b are described in 2.1.

Method B:

$$Q_L = \frac{\pi_c(1 + 2_g)}{\lambda_r \lambda_g (1 + g)} \frac{dl}{df},$$

where

c = velocity of light,
λ_r = free-space wavelength at resonance,
λ_g = guide wavelength at resonance,
g = normalized conductance of the tube,
dl/df = rate of change of the phase of the VSWR with frequency.

dl/df must be determined at the voltage maximum close to the plane of symmetry of the tube. Since it is usually not possible to make measurements at this position, a correction for the length of line must be made. This is

$$\frac{dl}{df} = \frac{dl'}{df} - \frac{n}{4} \frac{d\lambda g}{df},$$

where

dl'/df = measured slope of a line obtained by plotting the observed position of a voltage minimum as a function of frequency,
n = odd number of quarter wavelengths measured, at resonance, from the reference minimum to the place of symmetry of the tube.

The block diagram for Method B is the same as that shown in Fig. 2.

2.4 Mode Purity (ATR Tubes)

When the geometry of an ATR tube and its mount does not correspond to that of the waveguide transmission line, spurious modes can be excited at one or more points in the operating band. These cause an abnormally low VSWR to be set up by the tube at these frequencies, and, therefore, degrade the branching-loss performance of the duplexer. The presence of these spurious modes can be detected by measurement of the VSWR, as described below (see Fig. 4).

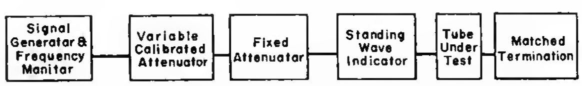

Fig. 4—Block diagram for mode-purity measurement.

The signal generator is connected to the standing-wave detector through a calibrated attenuator and a fixed attenuator of 10-db or greater insertion loss, the latter to isolate the signal generator from the high mismatch of the tube under test. The tube in its recommended mount is connected to the standing-wave detector and is followed by a matched termination. Standing-wave measurements[2] are made over the operating frequency range of the tube using the calibrated attenuator to determine the ratio, since commonly used standing-wave detectors usually do not conform to a known detection law over the wide variations of power level encountered.

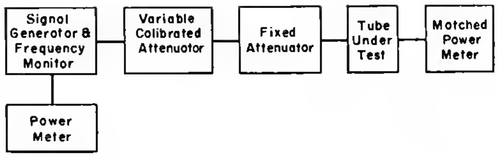

Fig. 5—Block diagram for insertion-loss measurement by Method A.

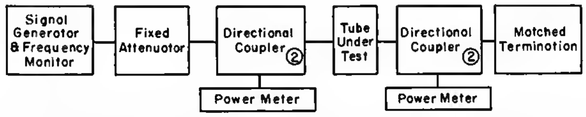

Fig. 6—Block diagram for insertion-loss measurement by Method C. (Two Directional Couplers may be of the 3- or 4-port type.)

2.5 *Insertion Loss* (*TR Tubes*)

TR insertion loss is a measure of the dissipative and reflective losses of the tube with the ignitor energized as specified and has, therefore, a direct bearing on the noise-figure of the receiver with which it is to be used. Three methods of determining TR insertion loss are described.

Method A:[1] (See Fig. 5.) The signal generator is connected to the tube under test through two attenuators in series; one has 10-db or greater attenuation to isolate the signal generator, and the other is variable and calibrated. The tube is followed by a matched detector. The insertion loss is measured by adjusting the calibrated attenuator to produce the same output of the detector when the tube is either in the circuit or is replaced by a waveguide of equal physical length. The change in the calibrated attenuator reading is the insertion loss, when the power output of the signal generator is constant throughout the test.

Method B: The insertion loss can be determined directly from the change in detected output when using essentially the same setup and procedure as in Method A, but with the calibrated attenuator omitted if the detector calibration is known. As in Method A, the power output of the signal generator must be held constant.

Method C: (See Fig. 6.) The signal generator is connected to the tube under test through an isolating pad of 10-db or greater attenuation and a directional-coupler-power-meter combination. The tube is followed by a similar power-measuring arrangement and a matched termination, as shown in the diagram. When the relative calibration of each directional-coupler-power-detector and power-meter combination is known to sufficient accuracy, the insertion loss of the tube can be determined. The directivity of the directional couplers must be high enough so that the power-meter readings will not be appreciably affected by the reflected power in the system.

Measurement of TR insertion loss should be made with the tubes placed in the mount specifically designed for the tube. Furthermore, care must be exercised in the use of any substitution-type measurements to insure that connectors in the test setup have sufficiently low VSWR as to cause negligible reflective losses; reflective losses due to the tube connectors must be included in the measured losses.

2.6 *Low-Level VSWR* (*TR Tubes*)

The low-level VSWR of a TR tube has a major effect on the noise figure of the receiver with which it is to be used. A mismatched tube may present an undesirable RF impedance to the receiver mixer. In addition, the mismatch causes reflective losses in the received signal. Two methods of determining VSWR are described.

Method A:[1] (See Fig. 7.) The signal generator is coupled to the standing-wave detector through a fixed attenuator having 10-db or greater attenuation to isolate the signal generator from the mismatch of the tube under test. The tube is inserted between the standing-wave detector and a matched termination in such a manner that the VSWR of the specified tube connectors are also properly included in the measured VSWR. Before the VSWR measurements are made, it should be established that the termination and any auxiliary connectors employed contribute only a negligible amount to the measured results. The ignitor should be energized as specified.

Method B: (See Fig. 8.) In this method the signal generator is an oscillator capable of being swept over the full operating frequency range of the tube under test. The generator is connected through an isolating attenuator to two directional couplers that sample the incident and the reflected energy. These couplers are followed by the tube under test and the matched termination, as shown. Assuming that the horizontal sweep of the oscilloscope is synchronous with the sweep frequency of the signal generator, and that the vertical deflection of the oscilloscope is properly calibrated, the entire VSWR-vs-frequency characteristic can be read directly on the oscilloscope screen.

As an alternative technique, a number of sequentially-pulsed fixed-frequency oscillators may be used in place of the sweep generator to provide a bar-graph presenta-

[2] SWR is commonly expressed as the change in the calibrated attenuator reading; *i.e.*, SWR $=20 \log_{10}$ VSWR.

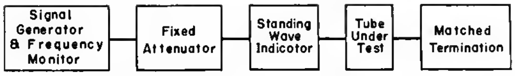

Fig. 7—Block diagram for VSWR measurement by Method A.

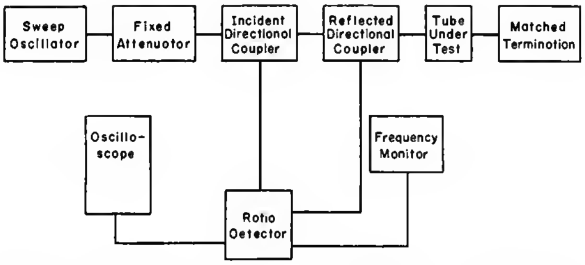

Fig. 8—Block diagram for VSWR measurement by Method B.

tion of the VSWR characteristic on the oscilloscope screen.

In addition to the precautions outlined in Method A, Method B requires that the directivity of the directional couplers exceed 40 db over the operating frequency band.

2.7 *Ignitor Interaction (TR Tubes)*

The loss introduced by the ignitor interaction is normally included in the over-all insertion-loss measurement, as described in 2.5. However, it is sometimes desirable to determine the loss due to the ignitor interaction alone in order to determine specific characteristics of the tube. To evaluate the loss due to the ignitor interaction, the power transmitted through the tube is measured with the ignitor off and then with the ignitor energized as specified. The ratio of these two powers expressed in db is the ignitor interaction (see Fig. 5).

2.8 *Low-Level Phase Shift (TR Tubes)*

For applications in which phase-comparison techniques are critical, the relative phase shift through one or more TR tubes must be maintained within specified limits over a given frequency range.

Method A:[1] (See Fig. 9.) The phase shift produced by a TR tube relative to a standard can be measured by feeding two signals derived from a common source into opposite ends of a standing-wave indicator. The interference of the two signals produces a high standing-wave pattern on the standing-wave indicator with sharp minima. When a TR tube is inserted into one of the lines in place of the standard, the phase shift of the tube causes a shift in the interference pattern. This phase shift is twice the measured shift in the position of a particular minimum.

It is advisable to adjust the amplitude of the two signals incident upon the standing-wave indicator to be equal so that the minima will be as sharp as possible.

Unless the phase characteristics of the attenuators in the ring circuit are known, so that appropriate compensations can be made for them, the attenuators should not be adjusted during the measurement.

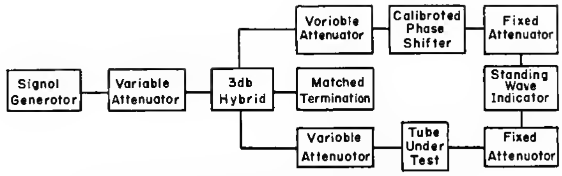

Fig. 9—Block diagram for low-level phase-shift measurement by Method A.

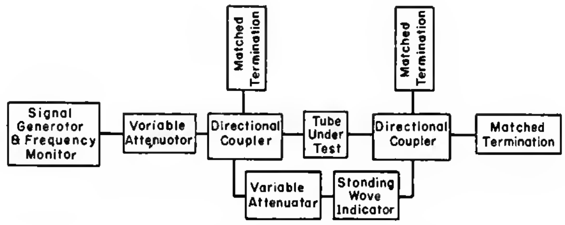

Fig. 10—Block diagram for low-level phase-shift measurement by Method B.

Method B: (See Fig. 10.) An alternative method of measurement uses the following circuit. The tube under test is placed between two directional couplers of approximately equal coupling coefficients. The loop is closed through a standing-wave indicator (Fig. 10) with an adjustable attenuator inserted in one of the two arms to compensate for any unbalance in the coupling coefficients of the two couplers. A standing-wave minimum is determined by inserting a standard in place of the tube under test. The shift in the minimum when the standard is replaced by the tube under test is one-half of the differential phase shift.

2.9 *Low-Level VSWR (Dual TR Tubes)*

Dual TR tubes are basically a combination of two band-pass tubes having a common gas fill, which are physically joined between two 3-db hybrid couplers to form a balanced duplexer. Although some of the characteristics of the dual tube can be adequately described in terms of the performance of each section, many duplexer parameters are dependent upon the relative characteristics of the two sections as well as on the associated hybrid couplers. Dual tubes are employed in many instances because in combination with well-designed hybrid junctions they provide lower duplexer branching loss, lower total leakage energy to the receiver, greater operating bandwidth, and eliminate the need for a separate ATR tube. Though the measurements outlined below are in many respects similar to those applicable to band-pass TR measurements, they are presented in order that those characteristics peculiar to dual tubes may be determined.

The performance of certain types of radar systems, such as monopulse and stacked-beam systems, may deteriorate due to excessive mismatch at the antenna terminals of the duplexers. The effect is one of decreased isolation between the several receiving channels. Since these radars may or may not employ isolators in the

transmitter circuit, two different conditions for low-level VSWR measurement are necessary.

Using the circuit of Fig. 11 indicated above, the measurement procedures and precautions are as outlined in 2.6. Where it is desired to simulate the condition of an isolator in the transmitter port, a matched termination is used. Where no isolator is used, the adjustable short circuit is set for maximum VSWR reading, thereby simulating a quiescent transmitting tube under the worst condition.

2.10 Transmitter-Receiver Isolation Dual TR Tubes

Certain types of transmitters are characterized by considerable interpulse noise output. Since some part of this noise appears at the receiver input terminals, the dynamic noise figure of systems employing these transmitters will be increased. A balanced duplexer can, within limits, alleviate this problem by maintaining some degree of isolation between the transmitter and receiver ports. The degree of isolation depends upon the combined performance of the hybrids and its dual TR tube.

Transmitter-to-receiver isolation is the ratio of the power incident upon the transmitter port to that available at the receiver port, expressed in decibels, with the other two ports terminated in matched terminations.

As shown in Fig. 12, the power levels at the transmitter and receiver ports are measured. The power incident upon the transmitter port may be measured through a calibrated directional coupler or by direct substitution of the power detector and power meter from the receiver port.

If the power detector and power meter from the receiver port are used, care must be taken to maintain the incident power at a constant level.

2.11 Loaded Q (High-Q TR Tubes)

The loaded Q of a TR tube is an indication of the operational bandwidth and to some extent an indication of the amount of image rejection to be expected in system operation.

Method A:[1] The tube is inserted, as shown in Fig. 13, between a matched generator and matched power detector and power meter and tuned to resonance at the specified frequency. The power meter is normally adjusted to some convenient indicator reading such as full-scale deflection. The generator is then detuned to frequencies above and below the resonance frequency for which the transmitted power is one-half the value at resonance. The difference between these two frequencies when divided into the resonance frequency gives the loaded Q_L, as defined by

$$Q_L = \frac{f_r}{\Delta f}.$$

Since the frequency interval to be measured is usually small, the use of a precision wavemeter or a heterodyne counter is required. In this way knowledge of the accuracy of the loaded-Q measurement may be determined. The power output should be maintained at a constant level over the frequency range of interest.

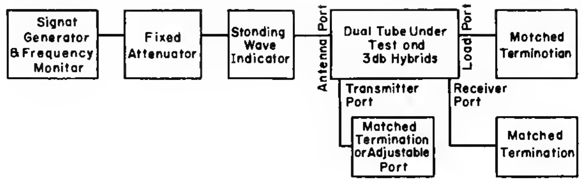

Fig. 11—Block diagram for low-level VSWR measurement.

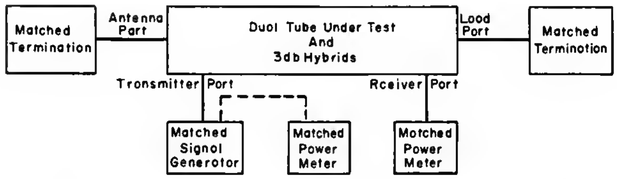

Fig. 12—Block diagram for transmitter-receiver isolation measurement.

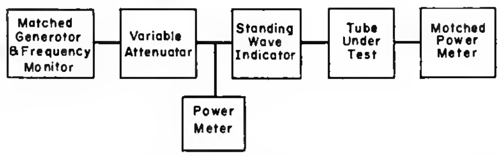

Fig. 13—Block diagram for loaded-Q measurement by Method A.

Method B: The loaded Q can also be measured by determining the input VSWR presented by the TR tube as a function of frequency. The half-power transmission frequencies can be related to the VSWR off resonance in terms of the VSWR at resonance. The half-power VSWR is then

$$S_{1/2} = \frac{S_r + 1 + \sqrt{S_r^2 + 1}}{S_r + 1 - \sqrt{S_r^2 + 1}}, \tag{1}$$

$$Q_L = \frac{f_r}{f_1 - f_2}, \tag{2}$$

where

S_r = VSWR at resonance,
$S_{1/2}$ = VSWR at half-power points,
f_1 and f_2 are those frequencies at which the VSWR is given by $S_{1/2}$ as in (1).

In each of the above cases it is assumed that the series losses connecting the cavity to the transmission line are not excessive. That is, the VSWR far from resonance is high (greater than 30 db).

2.12 Unloaded Q (High-Q TR Tubes)

A direct measurement of the unloaded Q, Q_u, of a cavity is impossible to perform because the mere process of connecting a transmission line to the cavity will load the cavity regardless of how loose the coupling may be. It is customary, therefore, to determine the un-

loaded Q from the loaded Q and some other parameter, usually VSWR, at resonance.

Referring to Fig. 13, the unloaded Q can be determined by measuring the transmission coefficient and the loaded Q.

$$Q_u = \frac{Q_L}{1 - \sqrt{T_r}}.$$

2.13 Resonance Frequency (High-Q TR Tubes)

Fixed-tuned cell-type TR tubes in a variety of external operational cavities are required to resonate over a variety of specified frequency bands. Tubes which resonate at specified frequences in the appropriate standardized cavities will be satisfactory in operational cavities.

The tube is mounted as indicated in Fig. 14, in the specified high-Q cavity, as shown in Fig. 15, and inserted between a matched signal generator and a matched power meter. The signal generator is varied to obtain a pronounced peak of the power meter. The signal-generator frequency at which the peak occurs is the resonance frequency of the TR tube in the specified cavity.

2.14 Tuning Range (High-Q TR Tubes)

The tuning range of a high-Q integral-cavity TR tube is the frequency band over which the tube can resonate. The frequency band of a superheterodyne radar receiver is generally determined by the tuning range of the TR tube and/or the local oscillator.

The tube is mounted in a transmission line between a matched generator and a matched power meter, as shown in Fig. 15. The tuning screw of the TR tube is rotated to one extreme position. The frequency of the generator is varied until the power meter indicates a sharp peak. The frequency of the oscillator at the point of maximum output of the power meter is the resonance frequency of the TR tube. The TR-tube tuner is then rotated to the other extreme position, and the resonance frequency is again determined. The tuning should be continuous and progressive, in which case the limit frequencies measured determine the band over which the tube is capable of being tuned.

Each of the matched attenuators should have an attenuation of at least 10 db.

2.15 Frequency-Temperature Drift (High-Q TR Tubes)

The resonance frequency of a high-Q TR tube will shift over the environmental temperature range encountered in operation. Excessive shift in the resonance frequency of the TR tube compared to the transmitter frequency will result in deterioration of the system sensitivity. The frequency-temperature drift is defined as the difference in the TR tube resonance frequencies at two specified environmental temperatures. Frequency-temperature drift by itself does not tell how much the resonance frequency changes under operation.

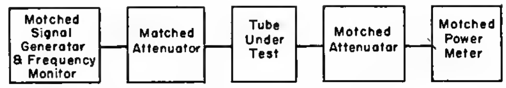

Fig. 14—Block diagram for determination of resonance frequency.

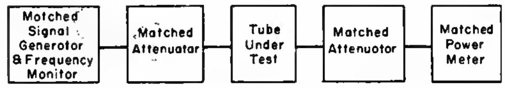

Fig. 15—Block diagram for tuning-range measurement.

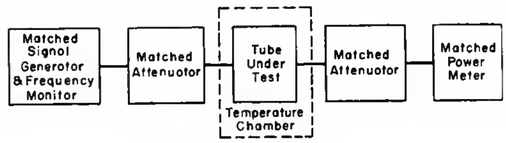

Fig. 16—Block diagram for measurement of frequency-temperature drift.

The resonance frequency for fixed-tuned cell-type TR tubes in external cavities is measured as indicated in 2.13.

In tunable high-Q TR tubes the magnitude of the frequency-temperature drift is strongly dependent upon the relative position of the tube cones. The frequency-temperature drift is, therefore, measured after the tube has been tuned to the desired frequency at room temperature. This is accomplished by inserting the TR tube between a matched signal generator and a matched power meter (see Fig. 16). With the signal generator set at the desired frequency, the TR tube is tuned to resonance as indicated by a pronounced peak on the output indicator. The frequency-temperature drift is then obtained in the same manner as that for a fixed-tuned cell-type TR tube, as previously indicated.

2.16 Insertion Loss (High-Q TR Tubes)

The insertion loss of an external-cavity cell-type TR tube has significance only when considered in association with a standardized cavity. When the combination is inserted into a transmission line between a matched generator and a matched power-measuring device, the insertion loss can be measured in the same manner as indicated in 2.5. In this case, the dummy tube does not consist of an equivalent length of line, but rather of one or more connectors that are as reflectionless and loss-free as possible.

The insertion loss of an integral-cavity TR tube can be measured as indicated in 2.5 with the tube tuned to the specified frequency.

3. High-Level Radio-Frequency Measurements

3.1 Recovery Time (ATR Tubes)

The recovery-time characteristic of ATR tubes can be an important factor in over-all radar-system performance. A substantial portion of the echo signal may be channeled into the transmitter rather than the receiver if the ATR tube is ionized during the receive por-

tion of the cycle. This may cause a deterioration of system performance with respect to its short-range, upper-angle coverage capabilities.

The ATR recovery time is the time interval required for the tube to deionize from the fully ionized condition to a point of 3-db duplexer insertion loss. It is assumed that the TR tube is lossless, and that the nonoscillating transmitter is mismatched to a VSWR of approximately 10 at its worst phase. This elapsed time is determined by the time interval between the end of the transmitter pulse and that at which the impedance of the ATR tube results in a 3-db loss. The 3-db loss is determined by measuring the impedance of the tube as a function of time. In a transmitter that has matched terminals the duplexer loss will be a maximum of 3 db, regardless of the ATR recovery characteristics.

The impedance of the ATR tube as a function of time is obtained by measuring the VSWR and the phase of the reflection coefficient using the test setup shown in Fig. 17.

The VSWR is measured as follows. The transmitter power is set to the specified value, and a low-level pulsed probing signal which can be varied in time with respect to the transmitter pulse is fed into the transmission line through a high-directivity coupler. The signal is detected by the probe in the standing-wave indicator and displayed on the scope. The probe is moved along the line for minimum deflection on the scope. The minimum is set to a convenient reference value. The probe is then moved to the maximum value, and the attenuator is then changed to reduce the deflection to the same reference value. The VSWR may then be determined by the change in the attenuator setting.

The phase is measured as follows. The position of the minimum of the voltage standing wave resulting from the probing signal is established as a phase reference point for the tube in the unfired condition. A shift in the position of the minimum occurs when the tube is ionized by the power from the transmitter. The shift in the position of the minimum is measured, and this distance, when converted to electrical degrees, gives the phase shift from the reference plane.

By varying the delay of this probing signal with respect to the transmitter pulse, and by measuring VSWR and phase at each time increment, the impedance of the ATR tube as a function of time from the end of the transmitter pulse may be determined by plotting the values on a Smith chart.

Precautions must be taken to prevent crystal burnout in the superheterodyne receiver. A TWT amplifier will act as a limiter to provide adequate protection.

3.2 *ATR Arc Loss*

The purpose of the arc-loss measurement is to determine the fraction of transmitter power absorbed within the ATR tube. It is an important measurement from two aspects: the power absorbed within the tube is a key factor in determining the maximum power at which the tube can be operated, and it is also a factor in determining the effective power transmitted by the system.

Method A:[1] One method of measuring arc loss is as shown in Fig. 18.

The transmitter power is set at a specified level, as indicated by the power meter. The signal is detected by the probe in the standing-wave indicator and displayed on the scope. The probe is moved along the line for minimum deflection on the scope. The minimum is set to a convenient reference value. The probe is then moved to the maximum value, and the attenuator is then changed to reduce the deflection to the same reference value. The VSWR may then be determined by the change in the attenuator setting.

The arc loss may be calculated from the relation

$$L = 20 \log_{10} \left(\frac{S+1}{S-1} \right).$$

Care must be taken to keep the incident power level below the breakdown point of the probe in the slotted section, and to mount the tube an integral number of half wavelengths from the main transmission line.

The same procedures can be used for measuring the arc loss of TR and pre-TR tubes.

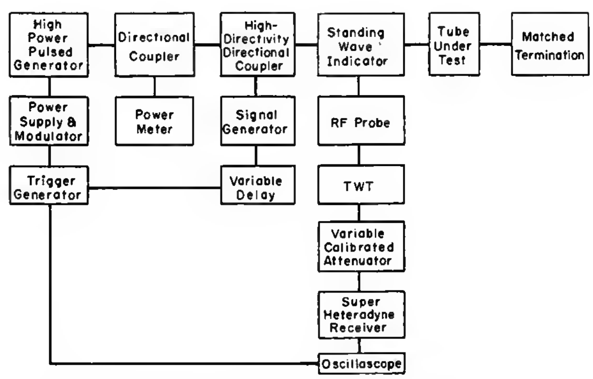

Fig. 17—Block diagram for ATR recovery-time measurement.

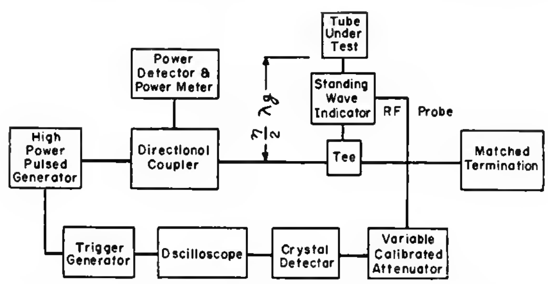

Fig. 18—Block diagram for arc-loss measurement by Method A.

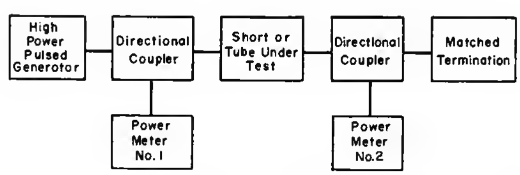

Fig. 19—Block diagram for arc-loss measurement by Method B.

Method B: The measurement can be made in the test setup shown in Fig. 19; this is a simple approximate method.

The tube under test is inserted in an appropriate mount. The transmitter power is set at a specified value, as indicated by power meter 1, and the power at the termination is noted, as indicated by power meter 2. The procedure is then repeated with a metallic short substituted for the tube under test. Arc loss in db is then defined as 10 log Ps/Pt, where Ps and Pt are the respective power readings at the termination with the short and with the tube in the line.

3.3 ATR High-Power-Level VSWR

The high-power-level VSWR is a measure of how well the tube is matched to the transmission line during the time it is in an ionized state. It can be measured as shown in Fig. 20.

The tube under test is inserted into an appropriate mount. The transmitter is set at the specified value, as indicated by the power meter, and the SWR is measured in the normal manner. Care must be taken to keep the incident-power level below the breakdown point of the probe in the slotted section.

The same procedure can be used for TR and pre-TR tubes.

3.4 ATR High-Power-Level Firing Time

The purpose of this test is to identify tubes that are slow leakers or contain an improper gas fill. It is a gross test in which a measurement is made (of the order of seconds) of the time required for the tube to ionize after the initial application of RF power when the tube has previously been in a quiescent state for a minimum of 24 hours.

The measurement can be made in the test setup shown in Fig. 21.

With a metallic short circuit in place of the tube, the RF power in the main transmission line is set at a specified level, which is generally below that of the operating power level of the tube. The metallic short circuit is replaced by the tube under test, and the time required for the tube to ionize, as indicated by a reading on power meter 2, is noted. An ionization time interval of less than 5 seconds is a rough indication that the gas fill is unchanged from the initial conditions. This test gives no indication of the interpulse firing time.

3.5 Recovery Time (TR and Pre-TR Tubes)

The measurement of recovery time entails the determination of the period from the cessation of the transmitter pulse to the time at which a signal through the TR tube is attenuated a specified amount. Under certain conditions the recovery time may limit the minimum-range capability of a radar system, since the amplitude of the return echo is dependent on the degree of deionization.

Method A:[1] The tube is inserted into a suitable TR-tube mount, usually a series Tee, with the high-power source set at specified operating conditions and the ignitor energized as specified. A simulated echo pulse is obtained from a low-level RF pulse-modulated signal source that is synchronized with the high-power transmitter and triggered by a variable-delay unit. Injection of the echo pulse into the main line is accomplished by a directional coupler which is placed between the TR tube mount and the transmitter, as in Fig. 22. The output signal is then mixed, amplified, detected, and placed on the vertical deflection plates of a synchroscope. The synchroscope horizontal sweep must be synchronized with the transmitter pulse. The low-level signal is varied in time with respect to the transmitted signal by adjustment of the delay unit. This method provides for a complete indication of the recovery period. The generally specified recovery point (3 db) is that point at which the echo signal is half the power level of the signal with the tube fully recovered.

A tube is considered to be fully recovered when the low-level signal transmitted through it is attenuated only by the insertion loss of the fully deionized tube. In most cases a tube is fully recovered prior to the next high-power pulse. However, in order to insure that the tube recovers fully during the interpulse time, the reference level should be checked with the transmitter off. The 3-db level can be measured on the synchroscope by comparing the amplitude of the attenuated signal with

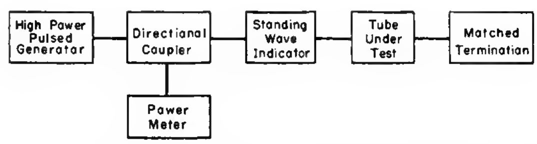

Fig. 20—Block diagram for high-power-level VSWR measurement.

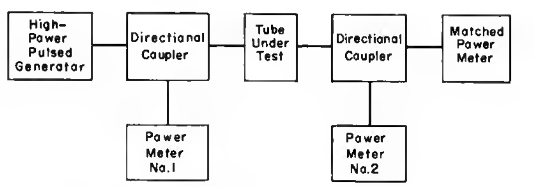

Fig. 21—Block diagram for high-power-level firing time.

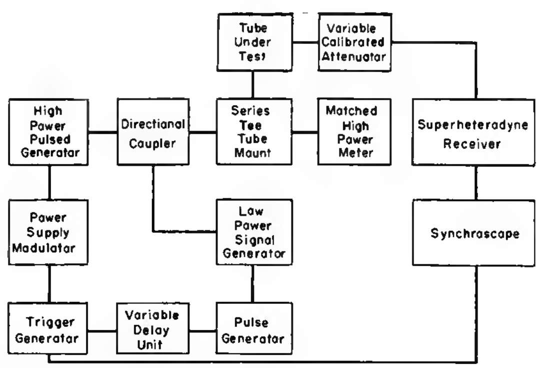

Fig. 22—Block diagram for recovery-time measurement.

the amplitude of the fully recovered signal.

Pre-TR tubes are tested in a similar manner, except that sufficient attenuation must be maintained in front of the crystal mixer to avoid crystal burnout.

In testing high-Q tunable TR tubes the echo-signal source should be set at the specified frequency for which the recovery time is to be measured. Since the resonance frequency changes during the recovery period, the cavity should be tuned to the echo-signal frequency with the transmitter off or at the end of the interpulse period. For accuracy of measurement the mixer must be matched. The recovery time is affected by a mismatched mixer which changes the loaded Q of the cavity and therefore the electron density of the tube.

The high-power source and the load should be well matched in order to prevent multiple reflections which can materially affect the recovery-time measurements. An indication of this variation can be obtained by insertion of a phase shifter between the high-power source and the TR tube.

Method B: Method A can be modified to permit a video technique of measurement. In such a case the pulse generator can be permitted to run freely. The superheterodyne receiver must be replaced by a crystal detector and a video amplifier. The synchroscope display will then indicate the complete recovery response. This technique is of value in determining very short recovery times. The echo-signal level should be kept low enough so that its magnitude does not contribute to increased recovery time.

3.6 *Position of Effective Short (TR and Pre-TR Tubes)*

The position of the effective short is the distance between some specified reference plane and the position of the equivalent short of the fired TR, ATR, or pre-TR tube. The position of the effective short should be controlled to prevent variations in the phase of the transmitted signal and excessive VSWR in the main line.

The position of the effective short is measured in units of length by determining the distance that a VSWR minimum point shifts when the fired tube replaces a physical short whose position is known with respect to the reference plane. A direct reading of the position of the effective short can also be obtained.

Method A:[1] With the equipment shown in Fig. 23, the RF power in the main line is set at a specified level. The position of the voltage standing-wave minimum is determined with the tube in the test circuit. The tube is then replaced by the metal shorting plate which established the reference position. The physical distance between the positions of the standing-wave minima of the tube and the reference short determines the position of effective short.

This technique is limited by the power-handling capability of the slotted section. The length of line between the Tee and the input window of the tube under test should be chosen for a good transmitter match.

Method B: The position of the effective short can also be determined with the arrangement shown in Fig. 24. The tube under test is inserted into one of the balanced arms of the magic Tee while the opposite arm is terminated in a calibrated adjustable short. With the generator output level adjusted to a specified value, the calibrated adjustable short is positioned for minimum reflected power as shown by a null on the power meter. The tube is then replaced by a reference short, and the adjustable calibrated short is repositioned until a null is again obtained at the power detector. The difference in the adjustable-short settings corresponds to the distance between the reference short and the short produced by the fired tube.

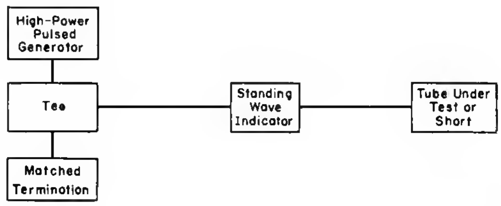

Fig. 23—Block diagram for measurement of effective-short position by Method A.

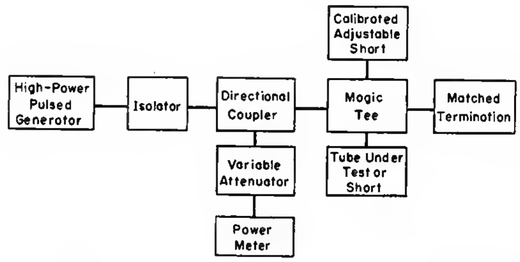

Fig. 24—Block diagram for position of effective-short measurement by Method B.

The calibrated adjustable short should be set before high power is applied so that there is approximately a quarter-wavelength difference in phase between it and the tube under test or reference short.

3.7 *Leakage Power (TR and Pre-TR Tubes)*

Leakage power is the RF power that passes through a fired tube and must be controlled in order to protect the receiver. For rectangular input RF pulses, the leakage pulse through the tube consists of two component parts: the spike portion of the pulse and the flat portion of the pulse. The term *leakage power* is also used to denote the total power in the leakage pulse. Measurements of leakage power are usually made with the tube mounted in series or shunt with the transmission line. The tube in its mount is inserted between a pulsed high-power level generator, usually a magnetron, and a matched high-power termination.

The formulas derived for all leakage-power parameters are based on the characteristics of the leakage power pulse. Therefore, all measurements of pulse dura-

tion refer to the leakage-power pulse itself. However, instrumentation problems often occur which make an accurate determination of the leakage-power pulse duration difficult. In that event reasonable accuracy may be obtained by measuring the transmitter pulse duration.

3.7.1 Leakage Power (Total Leakage): This method is used in cases where separation of the two components of the leakage pulse is not significant, as in high-Q TR tubes and pre-TR tubes. (See Fig. 25.)

The tube, in its mount, is placed in series or shunt with the transmission line between the generator and termination. The high-level pulsed RF power is applied by the generator at the required level. The ignitor is energized as specified. The leakage power as determined by a matched power detector mounted on the output termination of the tube is observed on a power meter. Assuming a rectangular leakage pulse, the measured average power is converted to peak leakage power by

$$p = \frac{P}{D},$$

where

p = peak leakage power,
P = average leakage power,
D = duty factor (pulse duration $t_p \times$ recurrence frequency f_p).

3.7.2 Spike Leakage Energy: Spike leakage energy is the term applied to the initial portion of the leakage pulse which is of short duration (of the order of 0.005 μsec and 0.05 μsec for cell-type and band-pass tubes, respectively). It is the excess energy in the spike that will damage a receiver.

Method A:[1] This method is recommended for measurement of band-pass tubes in which the shape of the leakage pulse approximates that shown in Fig. 26.

Measurement of the leakage is accomplished by use of a pulse duration t_{p2} (Fig. 26) which may vary over a range for which there is no substantial increase in leakage power. For many common tube types, this range is 0.1 to 0.25 μsec. The spike leakage energy is then

$$W_s = \frac{10^7}{f_p} P,$$

where

W_s = spike leakage energy in ergs per pulse,
P = average leakage power in watts,
f_p = pulse-recurrence frequency in cycles per second.

Method B: Referring to Fig. 25, the high-power generator is adjusted to operate at either of two pulse durations which are 2 to 1 in proportion or greater. With the ignitor energized as specified, the leakage power is measured at one and then the other pulse duration. The spike leakage energy is

$$W_S = \frac{10^7}{f_p}\left[P_1 - \frac{\Delta P t_{p1}}{\Delta t_p}\right],$$

where

W_s = spike leakage energy in ergs per pulse,
P_1 = average leakage power at the longer pulse duration t_{p1} in watts,
ΔP = difference in average leakage power of the two pulse durations t_{p1} and t_{p2} in watts,
Δt_p = difference in pulse duration times $= t_{p1} - t_{p2}$,
f_p = pulse-recurrence frequency in cycles per second.

This method is most applicable for the idealized leakage pulse as shown in Fig. 27. In order to maintain reasonable accuracy, t_{p1} should be approximately 1 μsec.

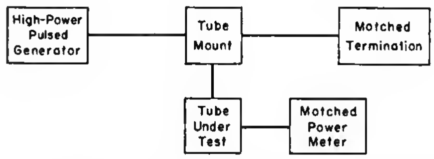

Fig. 25—Block diagram for leakage-power measurements.

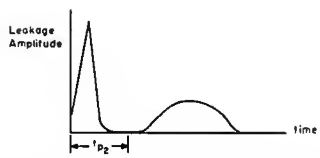

Fig. 26—Typical leakage pulse for band-pass TR tubes.

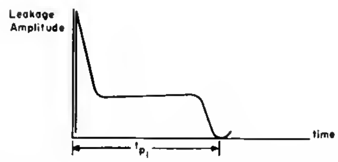

Fig. 27—Typical visual leakage pulse for cell-type tubes.

3.7.3 Flat Leakage Power: Flat leakage power is the term applied to the portion of the leakage pulse that occurs during a steady-state discharge. It is the power remaining when the spike leakage power is subtracted from the total leakage power.

Referring to Fig. 25, the high-power generator is adjusted to operate at either of two pulse durations which are in the proportion of 2 to 1 or greater. With the ignitor energized as specified, the leakage power is measured at one pulse duration, and then at the other pulse duration. The peak flat leakage power is obtained by use of the formula

$$P_f = \frac{(P_1 - P_2)}{f_p(t_{p1} - t_{p2})},$$

where

P_f = peak flat leakage power,
P_1 = average leakage power at the longer pulse duration,
P_2 = average leakage power at the shorter pulse duration,
t_{p_1} = longer pulse duration,
t_{p_2} = shorter pulse duration,
f_p = pulse-recurrence frequency.

In order to accurately determine the flat leakage power for band-pass tubes with the leakage envelope, as shown in Fig. 26, tp_2 should be long enough to include the beginning of the flat portion of the leakage pulse.

3.8 Minimum Operating Power (TR and Pre-TR Tubes)

The minimum operating power is generally dependent upon the particular system application. The high-power-level parameters of most general interest are the leakage characteristics. However, there are applications, for instance phase-sensitive systems, in which the minimum operating power will be determined by other characteristics such as arc loss, high-level VSWR, and the shift in the position of the effective short.

As the power level is reduced, the arc loss, high-level VSWR, position of the effective short, and leakage power vary, and degradation in the values of these parameters occur. The minimum operating power would be the point at which one of these parameters falls beyond a stated limit.

3.9 Phase-Recovery Time (TR and Pre-TR Tubes)

Phase-recovery time is the time required for a TR or pre-TR tube to deionize to the extent that the relative phase shift through the tube is reduced to a specified value. In certain applications where phase information is of importance, or in balanced duplexer arrangements, the introduced phase shift should be maintained within limits.

An arrangement for the measurement of the time required after the cessation of the transmitter pulse for a given signal to reach a specific phase shift in going through a tube is shown in Fig. 28. A synchronized low-power source provides RF pulses that can be transmitted through the tube at a specified time.

Initially, a phase reference is established by adjusting the variable calibrated phase shifter when the tube is unfired and the ignitor is energized as specified. Balance of the magic-Tee bridge is indicated by minimum signal at the power meter. With the high-power-level source on and the variable phase shifter set to a predetermined phase separation, the variable-delay unit is adjusted until a minimum signal is again obtained at the power meter. The reading on the variable-delay unit defines the time required for a given phase recovery of the tube.

In seeking the detector null the probing signal should be moved from the fully recovered toward the ionized

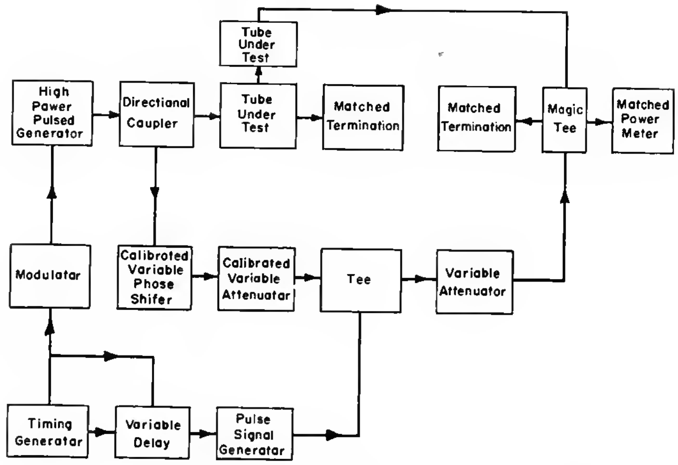

Fig. 28—Block diagram of phase recovery-time measurement.

condition of the tube in order to avoid phase-shift readings greater than 2π radians. If it is necessary to adjust the attenuator to sharpen the null indication, the phase characteristics of the attenuator must be known.

4. Ignitor-Electrode Measurements

All gas switching tubes which incorporate an ignitor electrode have a discharge created between that electrode and some portion of the tube structure. The ignitor electrode is customarily maintained at a negative potential with respect to the tube body so that the electrons released by the discharge will drift toward the adjacent gap. The ignitor structure thus serves to provide primary electrons to aid in the initiation of a high-power microwave gas discharge. In a TR tube this serves to reduce the spike leakage energy, and hence increases the receiver protection. In an ATR tube it insures firing on each transmitter pulse.

4.1 Ignitor Voltage Drop

The passage of a current through the ignitor electrode will result in a potential drop which, at a specified current, is called the ignitor voltage drop. The importance of the measurement of the ignitor voltage drop is to insure that the ignitor current is maintained within the required limits with a specified external circuit. In addition, its magnitude is a gross measure of the quality of the tube by providing an indication of serious degradation in the gas fill, or deterioration of the ignitor structure.

Method A:[1] In the circuit shown in Fig. 29 the supply is adjusted to obtain the specified ignitor current. The potential difference between the ignitor cap and ground as measured by a high-impedance voltmeter at least 20,000 ohms/volt, is the ignitor voltage drop.

Method B: An alternative method is to employ a voltmeter at the dotted position, in which case the ignitor voltage drop is $V_v - I_i R_s$, where V_v is the voltmeter reading, I_i is the ignitor current, and R_s is ohmic value of the series resistor. If high accuracy is to be maintained, R_s must be a precision resistor.

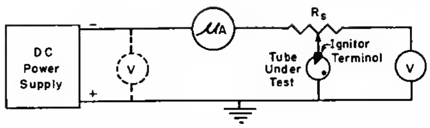

Fig. 29—Circuit diagram for ignitor voltage drop measurement.

In both methods a resistor should be placed as close as possible to the ignitor terminal to eliminate ignitor oscillations.

4.2 Ignitor Firing Time

A knowledge of ignitor firing time is important in radar operation. The ignitor must be fired before the initial application of high-level microwave power to assure the establishment of an RF discharge for receiver protection.

The following procedure will insure that the tube contains an approximately correct gas fill.

In the circuit shown in Fig. 29 the voltage and series resistance are adjusted to specified values without the tube in the circuit. The voltage is then applied to the tube. The time between the application of voltage and the initiation of an ignitor discharge, as indicated by the microammeter or other indicating device, is the ignitor firing time. This time is usually considerably less than five seconds.

4.3 Ignitor Oscillations

Under certain conditions, it is possible to obtain relaxation oscillations in the ignitor discharge. The supply of primary electrons would thus be varying and, during the period in which the ignitor discharge was extinguished, could be nonexistent. Thus the situation exists whereby excessive spike leakage is possible.

The mechanism by which relaxation oscillations occur can be explained with the aid of the following Figs. 30 and 31.

R is the first large current-limiting resistor in the circuit, C is the stray and lumped capacitance to ground between the ignitor discharge and the resistor R, and R_g is the effective resistance of the ignitor discharge. The voltage V_c across the ignitor will build up at a rate depending upon the RC time constant of the circuit until the breakdown voltage V_b is reached. The gap will

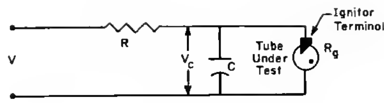

Fig. 30.

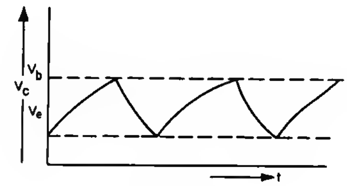

Fig. 31.

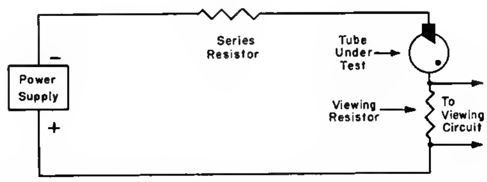

Fig. 32—Block diagram of ignitor oscillation measurement.

now break down and the capacitor will be discharged until the extinction voltage V_e is reached.

The discharge will then go out, and the cycle will be repeated. A required condition for this to occur is that R_g be small compared to the power-supply resistance, and that V be greater than V_b. A circuit for detecting ignitor oscillations is shown in Fig. 32. The viewing resistor has a value of 1 to 10 ohms, while the series resistor varies from 1 to 5 megohms to avoid keep-alive oscillations.

4.4 Ignitor-Leakage Resistance

The ignitor-leakage resistance is the resistance between the ignitor terminal and the tube body. It must be as high as possible so that the ignitor current is not shunted.

To determine the ignitor-leakage resistance, the resistance from the ignitor terminal to the tube body is measured by any method capable of determining resistances of the order of one hundred megohms or more. Care should be taken that the voltages used are not sufficient to produce an ignitor discharge.

Part 5

Phototubes

Subcommittee 7.4

Phototubes

M. Adelman, *Chairman* (1959–1962)
R. G. Stoudenheimer, *Past Chairman* (1955–1958)

S. F. Essig	D. H. Schaefer	A. H. Sommer
G. W. Iler	F. W. Schenkel	R. M. Sugarman
B. R. Linden		B. H. Vine

1. Introduction

A phototube is essentially a current generator with the output current a function of the total radiant flux on the photocathode.

1.1 Classification of Phototubes

Phototubes may be grouped into three general classes.

1) Vacuum-diode phototubes.
2) Gas-filled diode phototubes.
3) Multiplier phototubes.

For a given cathode, vacuum-diode phototubes have the least sensitivity and the greatest stability of the three classes.

Gas-filled phototubes are useful in sound reproduction from modulated light and other applications where the higher signal output is desirable but precise calibration is not required.

Multiplier phototubes have much greater sensitivity than the other two classes of phototubes. Very low light levels may be detected and the tubes may be used in low-impedance circuits.

1.2 Characteristics

Seven basic characteristics require specifications for common applications of phototubes.

1) Sensitivity (to various types of incident radiant energy).
2) Current amplification.
3) Current-voltage characteristics.
4) Dynamic characteristics (where they differ from static characteristics).
5) Electrode dark current.
6) Noise.
7) Peak output-current limitations.

Additional specifications are required in the field of scintillation counting.

Disturbing effects are frequently encountered, such as fatigue and nonuniform sensitivity over the cathode area. These effects are very important in many applications and make standard methods of testing and presenting the data desirable.

2. Sensitivity

In measuring sensitivity of a phototube a photocathode-envelope combination is considered as an integral unit. Sensitivity is thus expressed relative to radiation incident on the tube without correction for reflection and absorption losses of the envelope.

In all sensitivity measurements (except for illumination sensitivity) all radiation incident on the envelope should be directed toward the photocathode. The position and size of the illuminated area is important and should be stated if sensitivity is not uniform over the cathode area. The radiation should be essentially a parallel beam at normal incidence on the photocathode or at some stated angle with respect to a reference plane in the phototube.

Unless otherwise required for special purposes, the values of incident luminous or radiant flux and electrode voltages should be such that the output current is stable and a linear function of the flux.

Fatigue of the phototube should always be taken into account. If fatigue is observed in the particular tubes under test, measurements should be taken either after stabilization of response, or at a stated time after application of voltage and flux.

In all measurements of sensitivity dark current must be subtracted from output current. Ambient temperature should be stated.

2.1 Luminous Sensitivity

For these measurements a light source, an enclosure, and electrical circuits are needed as shown in Figs. 1 and 2.

The light source is commonly a tungsten-filament lamp employing either a flat ribbon filament or a concentrated coiled filament operated at 2870°K color temperature. The flat ribbon filament is usually more convenient because its light output in the direction normal to the plane of the filament is less critical with respect to small angular variations in orientation of the lamp. In either case, a projection-type lamp with a sufficiently heavy filament to assure a reasonable calibrated life is desirable. The bulb should be large enough

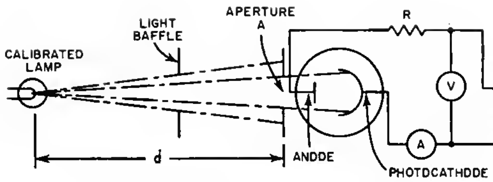

Fig. 1—System for measuring luminous sensitivity of diode phototubes.

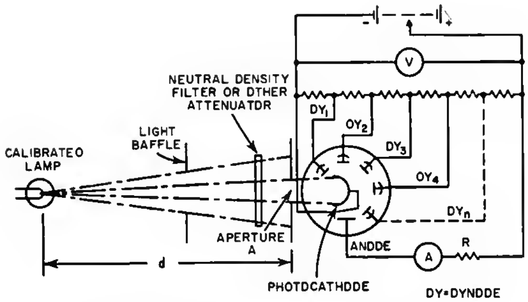

Fig. 2—System for measuring luminous sensitivity of multiplier diodes.

to allow the tungsten vapor from the filament to rise and deposit over the unused area of the glass. Before calibration, the lamp should be aged until the light output is sufficiently stable to meet the specific requirements. The lamp should be recalibrated periodically (after 10 to 50 hours of operation depending on type of lamp and precision desired) for light output and color temperature. In operation the lamp current must be accurately controlled. A variation of one per cent in lamp current may produce a variation in output current of six per cent in a blue-sensitive phototube.

The 2870° color temperature may be established by means of a lamp calibrated by a properly equipped laboratory.[1,2]

If frequent or prolonged measurements are to be made, it may be desirable to conserve the standard lamp by using a second lamp calibrated against the first by the following methods:

a) Red and blue filters are interposed between a phototube with S4 response[3] and the calibrated standard lamp. Filters should be selected to have widely separated pass bands within the useful range of the phototube spectral response and the spectral radiant energy distribution of the lamp. Currents are measured first with the blue and then with the red filters in place, and a "blue-to-red current ratio" is computed. The temperature of the second lamp is then adjusted until it produces the same ratio. Some suitable filters are:

	Red	Blue
Rohm and Haas (Plexiglas)	2444	2045
Corning	2404	5113 (one-half stock thickness)
	2408	
	2412	
	2418	

b) A visual color comparison of an uncalibrated lamp against a standard may also be made with a Lummer-Brodhun or similar photometer. Unless made by a skilled operator, visual comparisons may be subject to considerable error. A lamp may also be calibrated for luminance temperature by means of an accurate optical pyrometer. For a color temperature of 2870°K, tungsten has a luminance temperature of 2523°K.[4,5]

A phototube must be tested in an enclosure designed to prevent any extraneous light from falling on the phocathode. To prevent scattered light from the standard source from reaching the photocathode, a series of baffles with central openings having sharp edges should be placed between source and aperture. The aperture should be placed close to the tube and have a size chosen to expose a stated area of the cathode. A shutter may be incorporated for cutting off the light to measure electrode dark currents. Surface reflection from the baffles, aperture edges, shutter, and interior of the enclosure should be minimized by painting all surfaces a flat black. The amount of light falling on the phototube may be determined in two ways: 1) by calculation when a lamp of known horizontal candle-power is used, and 2) by measurement.

If calculations are to be used, the minimum distance from the aperture nearest to the phototube to the lamp filament must be great enough to permit the use of the inverse-square law approximation—at least ten times the maximum dimension of the filament. The value of luminous flux (F) entering the aperture and incident on the tube is

$$F = AI/d^2 \text{ lm},$$

where

A = area of the aperture,
I = intensity of the lamp, in candles, in the direction of the aperture,
d = distance from the aperture to the filament.

If measurements are to be used, an accurate light meter with a scale in footcandles should be employed. With the unobstructed sensitive surface of the meter in the plane of the aperture and facing the lamp, the value of luminous flux is

$$F = EA \text{ lm},$$

[1] The National Bureau of Standards will calibrate a lamp for color temperature.

[2] A. G. Worthing and D. Halliday, "Heat," John Wiley and Sons, Inc., New York, N. Y., pp. 466–469; 1948.

[3] Joint Electron Devices Engineering Council, "Spectral Sensitivity Characteristic of Phototube Having S-4 Response," Data Sheet J4-C3, April, 1949; and "Relative Spectral Response Curve Data," Data Sheet J4-C3-1, S-4, May 2, 1958.

[4] American Institute of Physics, "Temperature, its Measurement and Control in Science and Industry," Reinhold Publishing Corp., New York, N. Y., p. 1318; 1941.

[5] J. W. T. Walsh, "Photometry," Constable and Co., London, Eng.; 1953.

where

E = meter reading in footcandles,
A = area of the aperture in square feet.

Alternatively, a light meter may be calibrated to read, directly in lumens, the flux passing through the given aperture.

Luminous sensitivity: $S = I/F$ µa/lm, where

I = signal-output current in microamperes,
F = flux in lumens of 2870 K tungsten light.

2.1.1 Cathode-Luminous Sensitivity of a Phototube: For measurements of cathode luminous sensitivity, 0.1 lm is commonly used for a diode phototube and 0.01 for a multiplier phototube.

The circuit employed for measuring cathode-luminous sensitivity is shown in Fig. 1. When the the photocathode-luminous sensitivity of a multiplier phototube is measured, all electrodes except the photocathode are electrically connected and used as the anode. R is conventionally a 1-megohm resistor whose function is to limit the current through the tube, especially in the event of a glow discharge in a gas phototube. The ammeter may be a multiscale microammeter, and the value of voltage used will be the lowest value required to give a saturation photocurrent, generally 25 to 100 volts in vacuum tubes and 25 volts in gas tubes.

In general, since the relation between luminous flux and current at low light levels is linear for all types of phototubes, a measurement at one value of incident luminous flux should be adequate.

Cathode-luminous sensitivity (S_k) is

$$S_k = I_k/F \ \mu\text{a/lm},$$

where

I_k = cathode photocurrent in microamperes,
F = luminous flux in lumens of 2870°K tungsten light.

2.1.2 Luminous Sensitivity of a Phototube: The luminous sensitivity of a diode phototube is measured in the same manner as cathode-luminous sensitivity except that the rated, or otherwise stated, operating voltage is applied across the electrodes.

For measurements of luminous sensitivity of a multiplier phototube, a luminous flux in the range of 10^{-7} to 10^{-5} lm is commonly used. It may be necessary to use neutral-density filters to reduce the flux to these low values. A filter with a uniform transmittance throughout the visible and near infrared regions is most desirable. Since filter transmittance is generally not adequately uniform throughout this range, separate filter calibration is needed for tubes differing widely in their spectral sensitivity characteristics. Some available filters are:

a) Bausch and Lomb No. 31-34-38-01 (a set of four Inconel filters of nominal densities 0.3, 0.6, 0.9 and 1.2).
b) Wratten 96 (a series of thirteen standard densities).

Another means of reducing the light level is the use of a convex mirror system[6] as illustrated in Fig. 3. The luminous flux in lumens passing through aperture A may either be measured by using a calibrated multiplier phototube or calculated from the formula

$$F = AIr^2/(4D^2d^2)\left[1 + (\tfrac{1}{2}r\cos\phi)\left(\frac{1}{D} + \frac{1}{d}\right)\right] \cdot \left[1 + \left(\frac{r}{2\cos\phi}\right)\left(\frac{1}{D} + \frac{1}{d}\right)\right]1,$$

where

A = area of the aperture, and
I = intensity of the lamp, in candles, in the direction of the aperture.

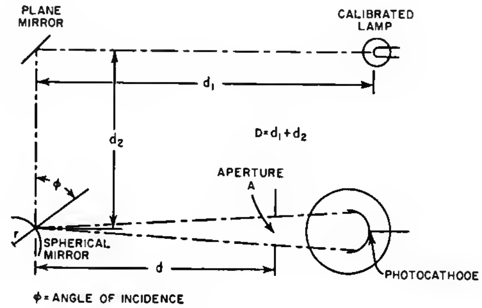

Fig. 3—Light attenuator system using a spherical and a plane mirror.

The reflection losses at both mirrors must be known and taken into account.

Another means of reducing light level fairly equally at visible wavelengths is to insert a colorless diffusing screen between the source and the aperture in the systems illustrated by Figs. 1 and 2. This screen is preferably a glass plate thoroughly grid-blasted on both sides, but may also be an opal glass. Maximum reduction will occur when the screen is placed approximately half-way between the light source and aperture A.

The circuit employed for measuring luminous sensitivity of a multiplier phototube is shown in Fig. 2. The ammeter may be a multiscale microammeter or milliammeter and electrode voltages used will have the values stated for the test.

Luminous sensitivity of a multiplier phototube is

$$S = I/F \ \text{a/lm},$$

where

I = signal-output current in amperes,
F = luminous flux in lumens of 2870°K tungsten light.

2.2 Illumination Sensitivity

For this measurement the same light sources, enclosures, and electrical circuits are used as for luminous sensitivity.

[6] R. W. Engstrom, "A luminous microflux standard," *Rev. Sci. Instr.*, vol. 26, pp. 622–623; June, 1955.

Instead of exposing the photocathode through an aperture to a given amount of luminous flux, the entire photocathode is exposed to a uniform illuminance (flux density). The illuminance to which the photocathode is exposed may be determined by one of the methods described in Section 2.1.

a) If the calibration constants of the lamp and geometry of the system are known, the illuminance is

$$E = I/d^2 \text{ fc},$$

where

I = horizontal intensity of the lamp, in candles,
d = the horizontal distance between the lamp and the photocathode, in feet.

b) Illuminance may be measured directly with a foot-candle meter placed in the position of the photosensitive surface of the phototube, and with its unobstructed face perpendicular to the light beam.

It is necessary that the light for this measurement flood the photocathode under stated conditions, usually at normal incidence to the photocathode or at some stated angle with respect to a reference plane of the tube.

The illumination sensitivity is

$$S_I = I/E \text{ a/fc},$$

where

I = signal-output current in amperes,
E = illuminance, in footcandles, of 2870°K tungsten light on the photocathode

2.3 Response to Filtered Light

For this measurement the same light sources, enclosures, and electrical circuits are used as for luminous sensitivity, but an appropriate filter is inserted between aperture A and the phototube. Photocathode current is measured. The response to filtered light may then be expressed as a current output for a stated filter and a stated value of luminous flux which would be incident on the tube if the filter were removed.

The amount of unfiltered light recommended is 0.1 lm. Suitable filters are:

a) Infrared—Corning code 2540, polished, stock thickness, or equivalent.
b) Red—Corning code 2418, polished, stock thickness, or equivalent.
c) Blue—Corning code 5113, polished, one-half stock thickness, or equivalent.

The ratio of photocathode current, with filter in place, to photocathode current, without filter, expressed as a percentage is frequently of interest.

2.4 Radiant Sensitivity (Monochromatic)

The required apparatus for measuring the radiant sensitivity includes a source of monochromatic radiation at various wavelengths throughout the spectrum and a means of measuring the radiant power incident on the phototube. The degree of spectral purity required is dependent on the rate of change of radiant sensitivity as a function of wavelength. Normally, spectral bandwidths of 100 A or less are sufficiently narrow. Precautions must be taken to insure that extraneous radiation does not reach the device under test.

A monochromator, such as the double-prism or double-pass variety, is needed to provide high spectral purity. Even with these instruments, filters may be required to eliminate scattered radiation at wavelengths where the source emits strongly, or the device under test has very high sensitivity. The primary radiation source will depend on the wavelength desired. In the ultraviolet range a mercury, xenon, or hydrogen discharge lamp is generally used. For the visible and near-infrared range, the tungsten incandescent lamp is satisfactory.

A calibrated[7] blackened-target thermocouple or thermopile plus the associated measuring circuitry are generally used for measurement of the exit radiant flux from the monochromator. In some instruments the thermocouple is an integral part of the monochromator, with the provision for shifting the beam of radiant energy emerging from the exit slit either to the thermocouple or through an exit window onto the photosurface. The monochromator should also include a light chopper so that the thermocouple drift, caused by changes in ambient temperature, and phototube dark current will have reduced effects on the measurement.

The electrical circuit for any phototube operated as a diode is the same as that shown in Fig. 1, and, for multiplier phototubes, the same as that shown in Fig. 2. The currents involved are frequently so small that the current meter must be a galvanometer of high sensitivity or a meter coupled to an amplifier. If the light is not chopped, a dc amplifier must be used, but chopped light and an ac amplifier are generally preferable.

2.4.1 Cathode-Radiant Sensitivity (Monochromatic) of a Phototube: This characteristic may be measured directly (at some stated wavelength) with the equipment described in Section 2.4 by operating the tube as a diode (see 2.1.1). The monochromatic cathode-radiant sensitivity is the quotient of the photocathode emission current by the value of incident radiant flux as measured by the thermocouple.

If the thermocouple is not calibrated, a relative spectral sensitivity characteristic may be obtained by the method described in 2.6. The absolute value of radiant sensitivity at the wavelength of maximum response can be obtained from the relative spectral-sensitivity characteristic and the luminous sensitivity by using the fol-

[7] R. Stair and R. G. Johnston, "Effects of recent knowledge of atomic constants and humidity on the calibration of the National Bureau of Standards thermal-radiation standards," *J. Res. NBS*, vol. 53, pp. 211–214; October, 1954.

lowing relation:[8]

$$\sigma = \frac{680 S_k \int \bar{y}_\lambda W_\lambda d\lambda}{\int R_\lambda W_\lambda d\lambda},$$

where

σ = cathode-radiant sensitivity at wavelength of maximum phototube response,
S_k = cathode-luminous sensitivity,
$\bar{y}_\lambda$ = relative sensitivity of the eye at wavelength λ,
W_λ = relative radiation from a tungsten lamp at 2870°K color temperature at wavelength λ,
R_λ = relative spectral sensitivity of the phototube at wavelength λ.

At any other wavelength the monochromatic radiant sensitivity $= \sigma R_\lambda$.

2.4.2 Radiant Sensitivity (Monochromatic) of a Phototube (Anode Radiant Sensitivity): For vacuum-diode phototubes, conditions for this measurement are identical with 2.4.1. For gas-filled and multiplier phototubes the output current is measured at the output electrode with stated operating voltages applied to all electrodes. All other test conditions described in 2.4.1 are applicable if phototube sensitivity is substituted for photocathode sensitivity where the latter appears.

2.5 Quantum Efficiency

Quantum efficiency at a stated wavelength is most conveniently computed from the monochromatic cathode-radiant sensitivity at the same wavelength by means of the standard formula

Quantum efficiency at wavelength

$$\lambda = \frac{1.24 \times 10^4 S_\lambda}{\lambda},$$

where

S_λ = cathode-radiant sensitivity in microamperes/microwatts at wavelength λ,
λ = wavelength of incident radiation in angstroms.

2.6 Spectral-Sensitivity Characteristic

The spectral-sensitivity characteristic of a phototube normally takes the form of a graph in which relative or absolute monochromatic radiant sensitivity is expressed as a function of wavelength. When the monochromatic radiant sensitivity is expressed in relative units, it is common practice to normalize the sensitivity to unity at the wavelength where the maximum sensitivity occurs.

[8] R. W. Engstrom, "Calculation of radiant photoelectric sensitivity from luminous sensitivity," *RCA Rev.*, vol. 16, pp. 116–121; March, 1955.

2.7 Uniformity of Sensitivity of Phototubes

In any phototube sensitivity over the cathode area will not be absolutely uniform because of nonuniformity of the cathode sensitivity (2.7.1) and other effects such as variation in electron collection efficiency over the cathode area (8.1). The uniformity may be expressed as percentage variation from the median sensitivity or the minimum-to-maximum ratio, or by a map of the cathode showing equal-sensitivity contour lines.

The uniformity of sensitivity may be measured by systematically scanning the cathode area with a small spot of light of constant flux and stated dimension while measuring the current from the tube (operated under stated conditions). The method of scanning may be manual, electromechanical, or electronic. For rapid qualitative inspection, a two-dimensional flying-spot scanning system may be utilized, and the electrical output of the tube presented on a cathode-ray tube. Variation in sensitivity over the scanned cathode area may be observed as shading in the presented picture. For taking quantitative data with this system, the ouput from a single-line sweep across the photocathode may be presented on a calibrated oscilloscope.

2.7.1 Uniformity of Sensitivity of a Photocathode: Measurements are the same as in 2.7, except that all electrodes (except the cathode) are connected together and serve as the anode (collecting electrodes). Sufficient voltage is applied to the collecting electrodes to ensure that all electrons emitted from the illuminated spot, regardless of its location in the cathode area, are collected.

2.8 Fatigue

Some phototubes under normal operating conditions exhibit a temporary change in sensitivity, termed fatigue. Fatigue may be expressed in terms of percentage change in sensitivity under stated operating conditions including all electrode voltages, total radiant flux, time of exposure to radiation, and output current. Sensitivity is measured at the beginning and the end of the operating period according to one of the methods specified in Section 2.1. Frequently a fatigue characteristic is obtained by measuring relative sensitivity as a function of time duration of the operating conditions.

3. Current Amplification

3.1 Gas-Amplification Factor of a Phototube

In either a gas or a vacuum phototube, the gas-amplification factor is the ratio of photocurrents for a stated luminous flux at two different anode voltages. For a simple vacuum phototube, 250 and 25 volts are commonly used; for a gas phototube, the operating voltage (usually 90 volts) and 25 volts. Normally this current ratio is not more than 1.2 in vacuum phototubes and is not a true measure of gas amplification. However, a high ratio indicates the presence of excess gas. In vac-

uum phototubes this ratio is commonly called gas ratio instead of gas-amplification factor.

Note: The light flux for these measurements should be stated and usually should not exceed 0.1 lm to avoid instability.

3.2 Current Amplification of a Multiplier Phototube

3.2.1 Over-All Current Amplification of a Multiplier Phototube: The current amplification is the ratio of the anode signal current to the cathode signal current at stated electrode voltages. For large values, it is difficult to make this measurement in one step because either the cathode signal current has to be made extremely low or the anode current will exceed the stated maximum (see 9.1). The amplification can be measured in steps by either of the two following methods.

a) The signal output current is measured at a sufficiently low luminous flux so that the value of the output current is within the linear region and fatigue effects are avoided. In general, the output current should be below 0.1 ma. The luminous flux is then increased by a known factor and the cathode current is measured. The over-all amplification is the ratio of original signal output current to final cathode signal current, multiplied by the factor by which the luminous flux has been increased. The change in luminous flux may be accomplished by one of the methods described under 2.1.2.

b) The incident flux is chosen so low that the anode current can be measured at the stated voltages. Next, the dynode current is measured in the lowest stage where it can be conveniently measured. The final stages are disconnected to avoid overloading, and the flux is increased until the dynode current has increased by a known factor, say 100. If the cathode signal current is still too low to be measured, the same procedure is repeated. The over-all amplification is finally determined as the ratio of the original signal-output current to the final cathode signal current, multiplied by the factor(s) by which the dynode current has been increased in each step.

Caution: At least two dynodes beyond the one for which the current is being measured must have normal voltage applied.

3.2.1.1 Average amplification per stage: The average amplification per stage of a multiplier phototube with n dynodes is the nth root of the over-all amplification.

3.2.2 Amplification per Stage: The amplification (δ) of an individual stage is the ratio of the number of emitted secondary electrons from that stage to the number of emitted electrons from the preceding stage. In order to measure the amplification of a particular stage, common practice is to measure the current in this and all preceding stages, including the cathode. Assume a photocathode signal current I_k at the photocathode, a signal current I_{dy1} measured at dynode 1 terminal, I_{dy2} at dynode 2 terminal and I_{dyr} at dynode r terminal. The primary current at dynode 1 is I_k, the secondary-emission current at dynode 1 is $I_k\delta 1$, and the net current measured at dynode 1 terminal is

$$I_{dy1} = I_k\delta_1 - I_k$$

$$\delta_1 = \frac{I_k + I_{dy1}}{I_k}.$$

Correspondingly, the amplification of the nth dynode is

$$\delta_n = \frac{I_k + \sum_{r=1}^{r=n} I_{dyr}}{I_k + \sum_{r=1}^{r=n-1} I_{dyr}}$$

4. Current-Voltage Characteristics

4.1 Diode Phototube

The photocurrent is measured as a function of the voltage across the tube at a given luminous flux. If a load resistance is used, the values of the voltage at the tube electrodes should be measured or calculated. If the tube has appreciable electrical leakage, the leakage current must be substracted from the output current at each voltage.

A family of curves may be obtained by repeating measurements with various values of luminuous flux.

With a gas phototube, care should be taken at high voltages to avoid a glow discharge that may do permanent damage to the photocathode. To prevent an arc discharge that would do damage to both tube and measuring instruments a limiting resistance of at least 10,000 ohms should always be used in series with the tube.

4.2 Multiplier Phototube

4.2.1 Anode-Last Dynode Characteristic: The anode current is measured as a function of the voltage between the anode and the last dynode while the other voltages on the tube are kept at a stated value. The characteristic is basically similar to that of a diode-vacuum phototube (see 4.1); however, as a result of the current amplification in the preceding stages, the anode current is frequently so large that at conventional voltages space charge may cause a nonlinear relationship between flux and anode current. Therefore, a family of curves of current vs voltage at constant flux should be obtained.

4.2.2 Over-All Amplification and Sensitivity Characteristics: These characteristics are measured with a stated ratio of electrode voltages by applying a variable over-all voltage to the divider supplying the voltages of the individual electrodes. Amplification and sensitivity increase so rapidly with over-all voltage that usually their logarithms are plotted against the voltage. Amplification may be measured as outlined in Section 3.2.1.

For the sensitivity characteristic, the test procedure outlined in Section 2.1 should be followed.

5. Dynamic Characteristics

5.1 Dynamic-Sensitivity Characteristics of a Gas Phototube

The dynamic sensitivity is of special interest for gas phototubes because it usually drops rapidly with a modulation frequency above a few thousand cps. Modulated light may be produced by various means, such as a rotating disc, a sound track on film, an electro-optical light modulator, or by a properly modulated cathode-ray tube with P15 or P16 very short persistence phosphor.[9]

Fig. 4 illustrates a rotating-disc light modulator and the electrical connections that may be used for this test. This rotating-disc system is accurate, convenient, and with properly adjusted optical system provides a stationary illuminated area upon the photocathode of the tube. The apertures are shaped so that the light is sinusoidally modulated. For example, if the fixed aperture looks like the half-cycle of a sine wave, the moving apertures should look like rectangles.

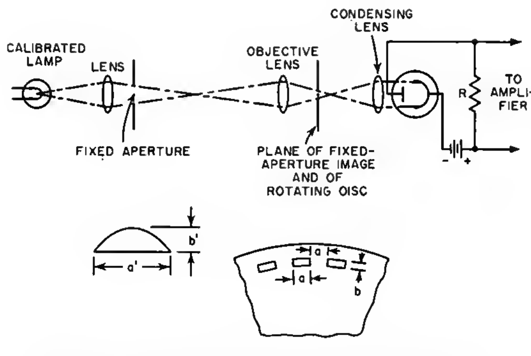

Fig. 4—Dynamic-sensitivity test arrangement.

The dimensions of the apertures should be chosen so that the dimensions of the image of the stationary aperture at the plane of the rotating disc are equal to those of the apertures of the rotating disc. The image of a' should equal a and the image of b' should not be larger than b. Uniform illumination is essential over the image on the rotating disc. The modulation frequency may be varied by varying the speed of rotation. For a given speed range, the frequency range can be extended by employing several sets of apertures in the rotating disc, proper fixed apertures being used for each set.

The ac output is measured as a function of frequency by a calibrated ac voltmeter connected across a resistance of about one megohm. The voltmeter impedance must be considered at the higher frequencies. The dynamic sensitivity at high frequencies is then determined relative to that at a fixed low frequency.

5.2 Pulse Response

Multiplier phototubes are frequently employed to observe light pulses that involve only a few photons. However, the methods employed in multiplier-phototube testing are based on the use of relatively large signals, such that the signal, either light or electrical, may be considered to be free of statistical variations. The signal may then conveniently be represented by either a step or delta function having a rise time negligible compared with the transit-time spread of the multiplier phototube. The pulse response of a phototube is ideally measured by using a delta function of light. If other input functions are used, the response should be corrected to that of a delta function; *e.g.*, if a step function is used, the output pulse is differentiated and the differentiated pulse is the equivalent output pulse for a delta input pulse. In measuring the output of a multiplier phototube, space-charge saturation or other sources of nonlinearity must be avoided. The spectral, spatial, and temporal distributions of the light incident on the photocathode must be stated.

In addition to the usual circuitry, a light-pulse generator and a detector capable of determining the shape and time of occurrence of the output pulse are required. Suitable light-pulse generators are 1) a mechanical light pulser,[10] 2) a spark discharge in hydrogen,[11] 3) a spark discharge in a mercury switch capsule.[12]

Caution: The glass capsule is filled with gas at high pressure. The capsule should be handled with great care to avoid injury from explosion. The mercury switch capsule is a very useful source for testing multiplier phototubes because it provides a short-rise-time and an accurately-synchronized electrical pulse. The detector is usually an oscilloscope, which must have sufficient bandwidth and sensitivity to prevent time and amplitude distortion of the output pulse. In many measurements, especially measurements of transit-time spread and pulse rise time, special viewing devices[13,14] and time coincidence circuits[15,16] are necessary.

5.2.1 Transit Time of a Multiplier Phototube: A suitable arrangement for measuring transit time is shown in Fig. 5. The length of the delay cable is adjusted until the time difference between the electrical marker pulse from the light pulser and a stated point on the multiplier-phototube output pulse can be determined on the

[9] Joint Electron Devices Engineering Council, "Description of Phosphors by Color and Persistence," Data Sheet J6-C3-1; October 1, 1959.

[10] J. B. Cladis, C. S. Jones, and K. A. Wickersheim, "High-speed light pulse shaper using a 5000 rps rotating mirror," *Rev. Sci. Instr.*, vol. 27, pp. 83–87; February, 1956.

[11] J. H. Malmberg, "Millimicrosecond duration light source," *Rev. Sci. Instr.*, vol. 28, pp. 1027–1029; December, 1959.

[12] Q. A. Kerns, F. A. Kirsten, and G. C. Cox, "Generator of nanosecond light pulses for phototube testing," *Rev. Sci. Instr.*, vol. 30, pp. 31–36; January, 1959.

[13] K. J. Germeshausen, S. Goldberg, and D. F. McDonald, "A high-sensibility cathode-ray tube for millimicrosecond transients," IRE Trans. on Electron Devices, vol. ED-4, pp. 152–158; April, 1957.

[14] R. Sugarman, "Sampling oscilloscope for statistically varying pulses," *Rev. Sci. Instr.*, vol. 28, pp. 933–938; November, 1957.

[15] Z. Bay, "Techniques and theory of fast coincidence experiments," IRE Trans. on Nuclear Science, vol. NS-3, pp. 12–28; November, 1956.

[16] I. A. D. Lewis and F. H. Wells, "Millimicrosecond Pulse Techniques," McGraw-Hill Book Co., Inc., New York, N. Y.; 1954.

oscilloscope.[17,18] If the light input and the marker pulses are both delta functions, the time delay is measured from the peak of the marker pulse to the peak of the phototube output pulse. If the light input is a step function, the equivalent time delay for a delta input function may be obtained by measuring from the peak of the marker pulse to the point of maximum slope of the output pulse.

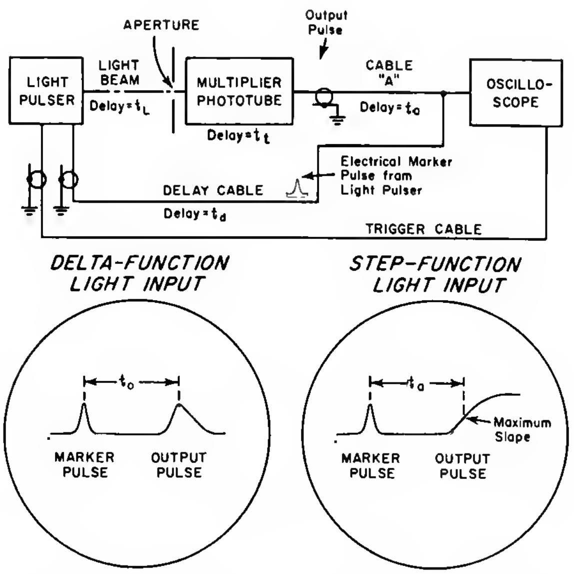

Fig. 5—System for measuring transit time and related characteristics. (Top—block diagram of equipment; bottom—oscilloscope traces.)

If the marker pulse precedes the tube output pulse by a time interval t_0 at the oscilloscope, if the electrical transit time of the delay cable and cable A are t_d and t_a, respectively, and if t_L is the required time for the light pulse to travel from the light source to the multiplier phototube, then the multiplier phototube transit time t_t is

$$t_t = t_0 + t_d - (t_L + t_a).$$

Since the transit time varies inversely as the square root of the applied voltages and is also a function of the position of the illuminated cathode area, the operating voltages and manner of illumination must be stated.

5.2.2 Transit-Time Spread of a Multiplier Phototube: The minimum output pulsewidth of a multiplier phototube is limited by transit-time spread of the tube. The arrangement shown in Fig. 5 is used to measure the transit-time spread. For a delta input pulse (or with output pulse corrected for a delta input pulse), the transit-time spread of the multiplier phototube is the full width at the half maximum of the output pulse displayed by the oscilloscope. Because the transit time depends upon the position of the illuminated area of the photocathode, the transit-time spread is a maximum when the entire photocathode is illuminated. Transit-time-spread measurements may be desired with either the full cathode illuminated or with the illumination on a circular area of stated size centered on the photocathode.

The rise time of the output pulse (for a delta-input pulse) is sometimes important and may be obtained using the procedure described for transit-time spread. Using the mercury-switch light pulse generator and considering the light pulse to be a delta function, the rise time is obtained more accurately than the corrected transit-time spread. The error in measuring transit-time spread arises because the decay time of the output pulse depends very significantly on the comparatively long decay characteristic of the light source[12] as well as the decay characteristic of the tube.

5.3 Variation in Transit Time with Position of Illumination

Variation in transit time from different illuminated areas of the photocathode may be the major contributing cause of a large transit-time spread and is consequently a useful characteristic for predicting transit-time spread. In making this measurement the adjustable aperture (Fig. 5) should be closed to the point where the diameter of the illuminated spot is about 5 per cent of the cathode diameter. When this small area of illumination is shifted over the photocathode, a variation in transit time is observed. Transit time at the center of the tube is usually used as the reference position. The variation in transit time from that at the reference position is measured as a function of both radius and azimuth. All electrode voltages should be stated.

6. Electrode Dark Current

The following test conditions are applicable to all dark-current measurements unless otherwise specified.

a) During measurements, the phototube should be placed in a light-tight box, as described in 2.1. For some phototubes, exposure to light within an hour immediately preceding the test may result in a higher dark current than for unexposed tubes.
b) The electronic circuitry is the same as that shown in Fig. 1 or 2.
c) Electrode voltages and ambient temperature should be carefully controlled at stated values because they affect the dark current of phototubes, especially of multiplier phototubes.
d) Socket leakage should be subtracted from electrode dark current to obtain the net dark current. This leakage varies with changes in ambient temperature, humidity, and voltage.

6.1 Anode Current

Anode dark current is measured at stated electrode

[17] Q. A. Kerns, "Improved Time Response in Scintillation Counting," IRE Trans. on Nuclear Science, vol. NS-3, pp. 114–120; November, 1956.

[18] R. V. Smith, "Photomultiplier transit-time measurements," IRE Trans. on Nuclear Science, vol. NS-3, pp. 120–122; November, 1956.

voltages, or at electrode voltages required to provide a stated luminous sensitivity. Possible causes of anode dark current are electrical leakage (tube, not socket), thermionic emission, field emission, residual gas ionization, and tube fluorescence. At low operating voltages, its major components are normally electrical leakage or thermionic emission, or both. Thermionic emission can be recognized by its temperature dependence. At high values of applied voltage and the dependent current amplification, the other dark-current components may become an appreciable part of the total dark current.

6.2 Equivalent Anode-Dark-Current Input

The equivalent anode-dark-current input of a phototube, in lumens, may be determined by measuring its anode dark current and luminous sensitivity at stated electrode voltages, or at electrode voltages providing a stated luminous sensitivity and making use of the following equation:

equivalent anode-dark-current input

$$= \frac{\text{anode dark current}}{\text{luminous sensitivity}}.$$

6.2.1 Equivalent Anode-Dark-Current Input vs Sensitivity Characteristic: This characteristic can be shown as a curve relating the values of the equivalent anode-dark-current (lumens) to values of luminous sensitivity (amperes per lumen).

6.3 Temperature Characteristic of Dark Current

A graph of anode dark current or equivalent anode-dark-current input as a function of equilibrium temperature normally covers the range of minus 65°C to the maximum rated-tube temperature. With certain cathodes erratic measurements may result from excessive cathode resistance when the tube is operated at low temperatures.

6.4 Electrode Dark Current in a Multiplier Phototube

High values of electrode dark current other than anode dark current may produce a serious loading effect, especially when the bleeder current is low, and thereby affect the voltage distribution to the electrodes. To measure the dark current of an electrode, the stated voltages should be applied to all electrodes and the current should be read on a meter inserted in the stated electrode lead.

7. Noise in Multiplier Phototubes

Fluctuations in the dark current limit the accuracy with which the instantaneous value of the signal output current may be determined, and fluctuations in the signal-output current limit the accuracy with which instantaneous values of light flux on the tube may be determined.

Noise is measured as a signal-to-noise ratio under stated operating conditions. In some cases an equivalent noise input is computed. Electrode voltages, incident flux, illuminated area, ambient temperature, and amplifier bandwidth should be stated. The light source is commonly a tungsten lamp operating at a color temperature of 2870°K. The incident flux should be modulated in a stated manner, preferably chopped, to provide rectangular light pulses with "on" period equal to "off" period. Noise measurements should preferably be made at low frequencies to include certain types of abnormal noise predominantly of low frequency.

As a typical example, the system shown in Fig. 6 may be used for noise measurements. Light from a tungsten lamp operating at a color temperature of 2870°K is chopped at 90 cps to provide rectangular light pulses with "on" period equal to "off" period. The output voltage across a load resistance R is passed through an ac amplifier and an RC filter (low-pass) having an equivalent noise pass band of 1000 cps. The db meter is calibrated to read the rms voltage across the load resistor in decibels above some reference voltage, usually 1 mv.

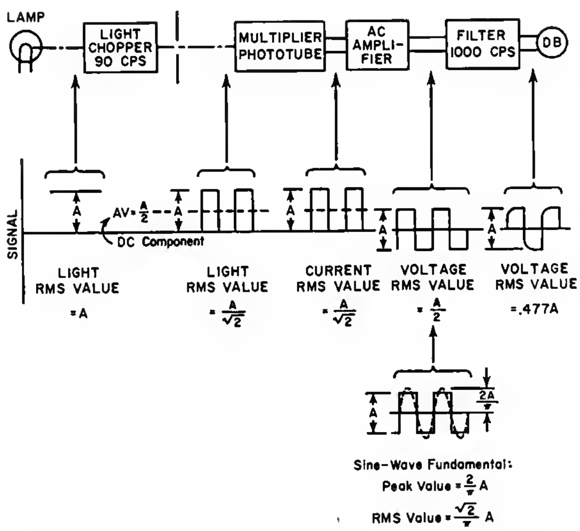

Fig. 6—Arrangement for signal-to-noise and related tests.

7.1 Signal-to-Noise Ratio

Under the stated operating conditions the rms (in some cases peak) signal current is measured at the output electrode with the light on. With the light off the rms dark-current noise is measured. The ratio

$$\frac{\text{rms (or peak) signal current}}{\text{rms dark-current noise}}$$

is the SNR. It is necessary to state whether peak or rms values of the signal are measured.

By use of the system shown in Fig. 6, the rms ac signal voltage (S in db) across the load resistor is read with the light on. The light is turned off and the rms noise voltage (N in db) across the load resistor is read. The rms SNR in db is $S-N$. S should be greater than N by 15 db if an error no greater than 2 db is desired. If the

measured SNR is less than 15 db, correction should be made for the presence of noise in the measured signal.

7.2 Equivalent Noise Input

The equivalent noise input for a bandwidth of 1 cps may be calculated using the SNR as measured in 7.1, with the system shown in Fig. 6. If the photocathode current during the "on" period is I_k amperes, then for a 1000-cps pass-band filter, the rms signal voltage (s in volts) across the load resistor is

$$s = 0.477\ \mu R I_k,$$

$$\mu = \text{multiplier current amplification},$$

$$R = \text{load resistance in ohms}.$$

For a bandwidth of one cps, centered at the chopping frequency, only the fundamental signal voltage would be measured.

The calculated rms signal voltage (s, in volts) of the fundamental is

$$s_1 = \frac{\sqrt{2}}{\pi}\ \mu R I_k \text{ (see Fig. 6)},$$

$$s_1 = 0.945s.$$

The calculated rms noise voltage (n_1 in volts) for a pass band of 1 cps, provided the noise spectrum is uniform, is

$$n_1 = \frac{n}{\sqrt{1000}},$$

n = measured rms noise (in volts) for pass band of 100 cps.

The equivalent noise input is

$$ENI = F\,\frac{n_1}{s_1},$$

where F is the luminous flux used in measuring the signal-to-noise voltage ratio (s/n), the values for s_1 and n_1, being introduced into the equation

$$ENI = \frac{F}{0.945\sqrt{1000}}\,\frac{n}{s}\ \text{lm}.$$

If the signal-to-noise ratio (S/N) is measured in db, where

$$20 \log s/n = S/N,$$

then

$$\log ENI = \log F - \log 0.945 - \frac{S/N}{20} - 3/2.$$

The noise and signal are both present in the signal measurement so that the luminous flux (F) should be adjusted to give a S/N ratio of no less than 15 db.

7.3 Ratio of Signal-to-Noise-in-Signal

The ratio of signal-to-noise-in-signal is dependent on the statistics of the electrons entering the multiplier structure. Low values of incident flux, low cathode sensitivity, and poor collection efficiency at the first few dynodes are factors which cause low signal-to-noise-in-signal ratios.

Under the stated operating conditions, the rms (in some cases peak) signal-output current is measured with the light on. If the light is not modulated, the dc output current may be used as the signal-output current. With the light on but not modulated, the rms ac noise is measured through a stated pass-band filter. The ratio

$$\frac{\text{rms (or peak) signal-output current}}{\text{rms noise-output current}}$$

is the signal-to-noise-in-signal ratio. It is necessary to state whether dc, peak ac, or rms values of signal current is measured.

Using the system shown in Fig. 6, the rms ac signal voltage (S in db) across the load resistor is read with the chopped light on. The chopper is then stopped and with a continuous light the rms noise voltage (N_s in db) across the load resistor is read. The signal-to-noise-in-signal ratio (S/N_s in db) is $S - N_s$.

8. Collection Efficiency

The collection efficiency[19] of the first dynode may be calculated from the signal-to-noise-in-signal ratio as measured in Section 7.3. If in the typical example described in Section 7, 10^{-7} lm peak (pulse maximum) is used for measuring the S/N_s,

$$S/N_s = 10 \log_{10} \epsilon S_k + 16.76,$$

or

$$\log_{10} \epsilon = \frac{S/N_s - 16.76}{10} - \log_{10} S_k,$$

where S_k is the measured cathode sensitivity in microamperes per lumen, and ϵ is the collection efficiency.

For an accurate measurement of collection efficiency, the test light flux should be adjusted to give a value of S/N_s of not less than 15 db. The constant at the end of the first equation is determined by test flux, waveshape at the output of the filter and frequency of chopping.

8.1 Uniformity of Collection Efficiency

Frequently different values of collection efficiency are measured as different areas of the cathode are illuminated with the test flux. The collection efficiency over the cathode may be plotted by means of contour lines of equal collection efficiency if the data for the plot are obtained by repeating the tests in Section 8.0 for a large number of points over the cathode, using for each measurement of light a small spot whose diameter is about 5 per cent of the photocathode diameter. Fre-

[19] R. W. Engstrom, R. G. Stoudenheimer, and A. M. Glover, "Production testing of multiplier phototubes," *Nucleonics*, vol. 10, pp. 58–62; April, 1952.

quently the data can be obtained using a more rapid, though less accurate, method by measuring the collection efficiency at the point on the cathode where the greatest signal output is obtained. Relative sensitivity is measured at all other points on the cathode by the method described in Section 2.8. Relative sensitivity multiplied by peak collection efficiency gives a close approximation to the collection efficiency over the entire cathode if cathode sensitivity is relatively uniform. If cathode sensitivity varies, the relative over-all amplification may be calculated for each point of illumination on the cathode. If the relative amplification for each point is expressed as a per cent of the maximum amplification, then the collection efficiency for each point on the cathode may be obtained by multiplying relative amplification by the maximum collection efficiency.

9. Peak-Output-Current Limitations

The output current of a phototube can be limited by two effects: 1) space charge, and 2) high resistivity of the photocathode. In general, the first effect is likely to occur in the last stages of multiplier tubes where the currents may be relatively high, or in diode phototubes exposed to intense flashes of light, while the second effect is found only in phototubes with semitransparent photocathodes.

Space-charge saturation in a multiplier phototube can be distinguished from saturation caused by a high-resistivity cathode by measuring the output current with increased voltage between some of the early dynodes to increase the current amplification. The output current will be increased if the saturation is caused by cathode resistivity, but it will not be affected if the saturation is due to space charge in one of the last few dynodes. For vacuum-diode phototubes, the space-charge-saturated output current is proportional to the $\frac{3}{2}$ power of the anode voltage. Any wide deviation from this relationship indicates high cathode resistivity. Drift of the saturation output current or temperature dependence indicates high cathode resistivity.

9.1 Space-Charge-Limited Output Current

Space-charge limitation results in an anode current which increases less than linearly with illumination. In the extreme case of complete space-charge saturation, the anode current remains constant with increasing illumination.

It is necessary to measure the output current while the illumination is varied by known amounts; this is best done by the use of calibrated neutral filters or by changing the distance between light source and photocathode and applying the inverse-square law. As the deviation from linearity starts very gradually, an exact value for the permissible peak output current for linear operation can only be stated if the permissible deviation from linearity is defined. For most purposes, a deviation of 5 per cent is a reasonable value.

Output current may be measured by dc methods up to values at which electrode dissipation exceeds safe values or where fatigue becomes an important factor, and by pulse methods beyond these limits. For dc methods, a steady illumination is used and the current is measured with a dc meter. In pulse methods the incident light is pulsed. Peak values of incident light flux and output current are measured. The cathode-ray oscilloscope is commonly used as the current indicator.

Space-charge limitation in simple phototubes occurs only at very high levels of illumination and is usually not encountered, except where phototubes are exposed to intense flashes of light. Measurements of output current in the nonlinear or saturated range require a pulsed light source of high peak intensity.[20] If the source does not give repeated pulses of the same peak intensity, an auxiliary phototube operated in the linear range (by enclosing it behind a neutral density filter) may be used to measure the relative peak intensities of the flashes. Absolute measurements of peak intensity are seldom required. Peak values of current are measured as the peak value of the incident light flux is varied. The output circuit should have a sufficiently wide pass band to avoid changing the shape of the output pulse. The anode voltage should be carefully adjusted to the proper stated value.

Space-charge limitation in multiplier phototubes may be measured by the dc method up to values where electrode dissipation exceeds safe values. Beyond this value, pulse methods are required, but the intensity of the source can be much lower than is used for simple phototubes. Suitable light sources are pulsed cathode-ray tubes with very-short-persistence phosphors such as P15 or P16,[8] gas-discharge lamps, or incandescent lamps with mechanical shutters. Otherwise, the method is the same as that outlined for simple phototubes. All electrode voltages should be carefully adjusted to the stated values.

9.2 Peak Output Current Limited by High Cathode Resistivity

Limiting cathode current values of 0.1 μa or less may be encountered in semitransparent cathodes of the antimony-cesium type and at somewhat higher values for other types of semitransparent cathodes. High cathode resistivity may result in an anode current which increases nonlinearly with illumination. The test for linearity is the same as that described under 9.1.

If a phototube with semitransparent cathode is to be used at low temperature (*e.g.*, liquid nitrogen), the linearity test must be carried out at this temperature because the resistivity usually increases with decreasing temperature. Temperature and electrode voltages should be adjusted to the proper stated values.

[20] W. C. Hall, B. M. Horton, J. W. Keller, and S. H. Liebson, "A High-Resolution, High Intensity Scintillation Detector," Naval Research Laboratory, Washington, D.C., NRL Rept. 3927; 1952.

10. Definitions

Equivalent Noise Input (of a Phototube). The value of incident luminous (or radiant) flux which, when modulated in a stated manner, produces an rms signal output current equal to the rms dark-current noise both in the same specified bandwidth (usually 1 cps).

Sensitivity (of a Photosensitive Electron Device). The quotient of the output quantity (such as current, voltage, etc.) by the incident radiant flux or flux density under stated conditions of irradiation.

Note 1: This term is general and the measured value depends on many factors such as spectral distribution of incident light, units used for measuring incident radiation, electrode under consideration, and output quantity. In several specific types of sensitivity the conditions of the test are more restricted. Where a well-defined specific sensitivity term is not applicable, care must be taken to state all the pertinent conditions under which sensitivity is measured.

Note 2: If the output has some value in the dark, the dark value is subtracted from the output when irradiated.

Sensitivity, Dynamic (of a Phototube). The quotient of the modulated component of the output current by the modulated component of the incident radiation at a stated frequency of modulation.

Note: Unless otherwise stated the modulation waveshape is sinusoidal.

Sensitivity, Radiant (Camera Tubes or Phototubes). The quotient of *signal output current* by incident radiant flux under specified conditions of irradiation.

Note 1: Radiant sensitivity is usually measured with a collimated beam at normal incidence.

Note 2: The incident radiant flux is usually monochromatic at a given wavelength. If the radiant flux is not monochromatic, its source must be described.

Transit Time (of a Multiplier Phototube). The time interval between the arrival of a delta function light pulse at the entrance window of the tube and the time at which the output pulse at the anode terminal reaches peak amplitude.

Transit-Time Spread. The time interval between the half-amplitude points of the output pulse at the anode terminal, arising from a delta function of light incident on the entrance window of the tube.

Part 6

Microwave Tubes

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Introduction

The methods of test described here are representative of current practice. They have been organized by three groups, the first concerned primarily with non-operating measurements, the second with microwave oscillators, and the third with microwave amplifiers. In many instances, it is neither possible nor advisable to recommend a single test or method of measurement for a given parameter because conditions as to magnitudes involved or physical environment may differ widely. Furthermore, accurate and reliable tests may be performed with a variety of test components. Recommendations have been made as to reliable methods that have been investigated carefully and are known to be widely used.

The scope of recommended test measurements, although not all-inclusive, does include those properties most important in the evaluation of microwave tubes. The determination of resonance frequency and phase, the measurement of impedance, the measurement of power and attenuation, and the determination of those phenomena which result in distorting the characteristics of tube outputs—for example, noise, spurious oscillations and intermodulation and the measurement of modulation characteristics—represent the major portion of these proposed standards.

Where possible, references to current or past literature have been listed to give support to the methods described and to provide the reader with additional detail and information. Precautions in the use of these tests have also been provided where needed, but it is assumed that the user of microwave test equipment is in general well versed in the art.

A. Nonoperating Characteristics

The purpose of this section is to describe the basic methods of measuring the nonoperating characteristics of vacuum microwave tubes. These measurements are useful in understanding and analyzing the circuit parameters and their relation to operating microwave tubes. They permit the determination of fundamental quantities, such as resonance frequencies and Q's. From these quantities, circuit efficiency and pulling figures may be calculated.

1. Resonance-Frequency Measurements

1.1 Reflection Method

A signal generator loosely coupled by means of an attenuator is connected to a transmission line including a frequency-measuring device and a standing-wave detector, as shown in Fig. 1. The transmission line is terminated by the microwave tube under test. The signal generator is tuned over the frequency range involved; the frequency at which minimum coefficient of reflection occurs is the resonance frequency.

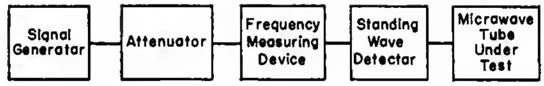


Fig. 1—Block diagram for reflection method of measuring resonance frequency.

1.2 Transmission Method

If the microwave tube has more than one coupling circuit, the frequency of resonance may be measured by the transmission method, as shown in Fig. 2. A signal loosely coupled by means of an attenuator is supplied to the microwave tube under test. The frequency at which the maximum percentage of power transmission through the microwave tube occurs, as indicated by the detector, is the resonance frequency.

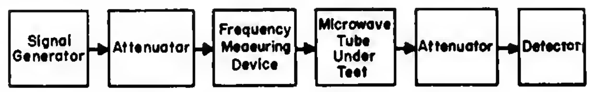


Fig. 2—Block diagram for transmission method.

2. Q Measurements

For the purpose of making microwave-tube measurements, Q's will be defined in terms of energy as follows:

$$Q = 2\pi \frac{\text{Energy Stored}}{\text{Energy Dissipated per Cycle}} .$$

The unloaded $Q(Q_U)$ takes into account only the dissipation resulting from internal losses within the microwave-tube resonant circuit. The external $Q(Q_E)$ takes into account only the external dissipation in the load when the output of the microwave tube is connected to a reflectionless transmission line. When more than one output circuit exists, each external $Q(Q_{E1}, Q_{E2}, \ldots)$ must be considered. If reflectionless transmission is assumed, Q_L (the loaded Q) is given by the following:

$$\frac{1}{Q_L} = \frac{1}{Q_U} + \sum \frac{1}{Q_{E_n}} .$$

For only one output

$$\frac{1}{Q_L} = \frac{1}{Q_U} + \frac{1}{Q_E} .$$

Various methods exist for determining these factors. These methods are indirect and are based on measurements of SWR and a position of standing-wave minimum as functions of frequency. In order to utilize admittance measurements to measure Q's the modes of the circuits must be well separated so that the equivalent circuit of Fig. 3 applies.

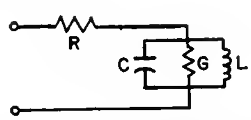


Fig. 3—Equivalent circuit of a single-output microwave tube if only one mode is considered.

It is possible to determine Q's by measurement of the SWR and minimum position of voltage minimum as functions of frequency, using equipment shown in Fig. 1.[1] The shift of minimum position as a function of frequency conforms to either Fig. 4(a) or (b), if mode separation is adequate.

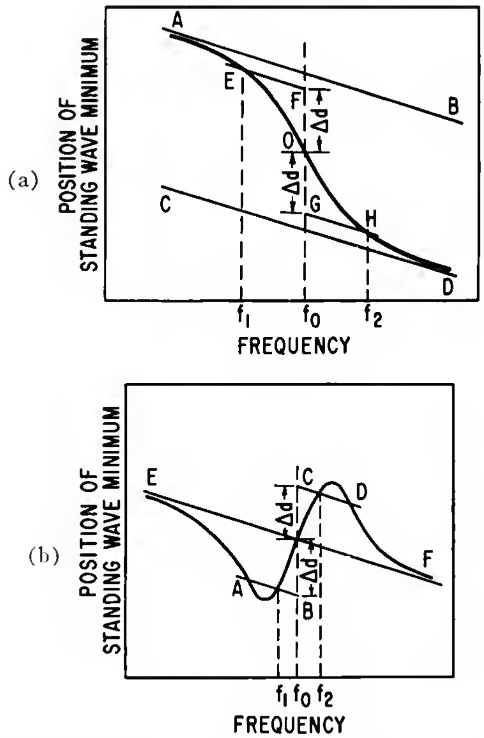


Fig. 4—Standing-wave minimum position as a function of frequency. (a) Overcoupled case. (b) Undercoupled case.

If the output circuit losses are small, the equivalent circuit of Fig. 3 is modified by neglecting the series resistance R. The lossy-output case has been considered by Slater[2] and by Malter and Brewer.[3] Other methods of measuring Q's have been described in the literature.[4,5]

2.1 *Overcoupled Case (Output Losses Neglected)*

The overcoupled case [Fig. 4(a)] is identified as the one in which the position of standing-wave minimum shifts by one-half guide wavelength as the frequency passes through resonance, and at resonance, the minimum position is shifting towards the microwave tube with increasing frequency.

In Fig. 4(a), AB, and CD represent the asymptotes that the plot of minimum position against frequency approaches far from resonance. Point O is the point on the curve that is midway between AB and CD. This point corresponds to the minimum value of SWR, and therefore to the resonance frequency, to be designated f_0.

The following expression is used to locate the half-power points E and H:

$$\frac{\Delta d}{\lambda_g} = \pm\left[\frac{1}{4\pi}\tan^{-1}\left(\frac{\rho_0+1}{\rho_0-1}\right)+\frac{1}{16}\right],$$

where ρ_0 is the voltage standing wave ratio at resonance. The ratio $\Delta d/\lambda_g$ as a function of ρ_0 is shown in Fig. 5(a). Distance Δd corresponds to OF and OG for Fig. 4(a). Points E and H can then be determined by projecting lines EF and GH from the f_0 line to the curve in such a manner as to take into account the change of electrical length of the transmission line with frequency. For small frequency differences, AB, CD, EF, and GH are approximately parallel. From points E and H, f_1 and f_2 are determined. It is then possible to compute the Q's:

$$Q_L = \frac{f_0}{f_1 - f_2}$$

$$Q_U = Q_L(1+\rho_0)$$

$$Q_E = Q_L\left(1+\frac{1}{\rho_0}\right).$$

Circuit efficiency is given by

$$\eta_c = \frac{Q_L}{Q_E} = \frac{\rho_0}{1+\rho_0}\cdot$$

Cold pulling figure p.f. is given by

$$\text{p.f.} = \frac{0.42 f_0}{Q_E}\cdot$$

This is a guide to the operating pulling figure,[6] which may be greater by 20 per cent or more than the cold pulling figure.

Instead of using the plot of standing-wave-minimum position as in Fig. 4(a), it is possible to determine Q by using the plot of VSWR as a function of frequency, as shown in Fig. 6. (It is necessary to identify the shift of standing-wave-minimum position as representing the over-coupled case.) The value of SWR at either half-power point is then given by the expression

$$\rho_1 = \rho_2 = \frac{\sqrt{2}+\sqrt{\left(\dfrac{\rho_0-1}{\rho_0+1}\right)^2+1}}{\sqrt{2}-\sqrt{\left(\dfrac{\rho_0-1}{\rho_0+1}\right)^2+1}}\cdot$$

[1] G. B. Collins, "Measurements of the resonant system," in "Microwave Magnetrons," McGraw-Hill Book Co., Inc., New York, N. Y., M.I.T. Rad. Lab. Ser., vol. 6, pp. 698–730; 1948.

[2] J. C. Slater, "Microwave Electronics," D. Van Nostrand Co. Inc., New York, N. Y., sect. 5.2, p. 89, *et seq.*; 1950.

[3] L. Malter and G. R. Brewer, "Microwave Q measurements in the presence of series losses," *J. Appl. Phys.*, vol. 20, pp. 918–925; October, 1949.

[4] E. D. Reed, "Sweep frequency method of Q measurement for single-ended resonators," *Proc. NEC*, vol. 7, pp. 162–172; 1951.

[5] M. Beverly, "Inflection point method of measuring Q at very high frequencies," *Sperry Engrg. Rev.*, vol, 4, p. 16; May–June, 1951.

[6] Collins, *op. cit.*, pp. 324–326, 734.

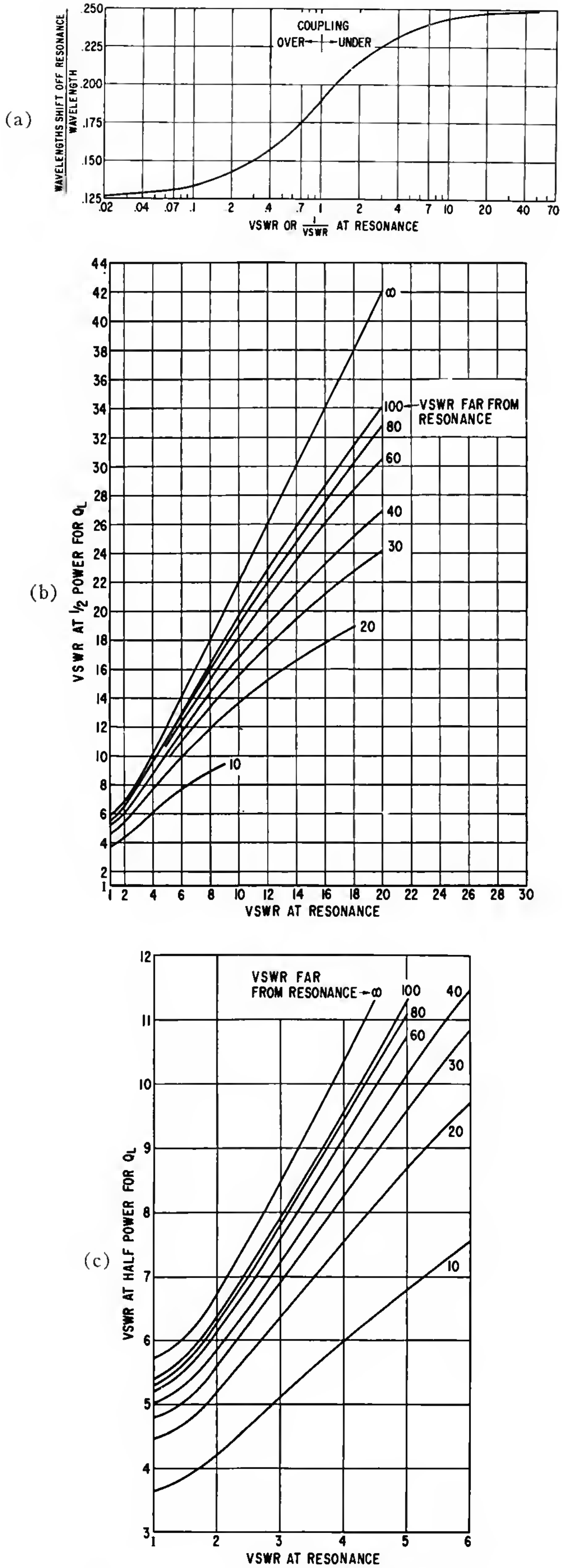


Fig. 5—Curves for determining half-power points from standing-wave measurements.

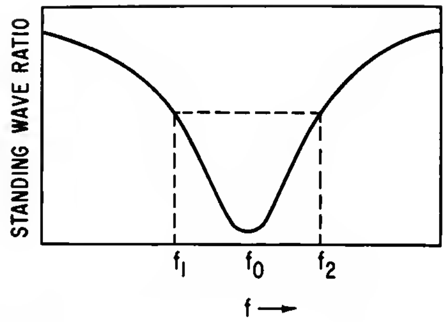


Fig. 6—Voltage of SWR as a function of frequency.

When ρ_1 and ρ_2 are established, f_1 and f_2 may be determined from the curve, and Q's calculated as before. It is convenient to use a plot of ρ_1 or ρ_2 as a function of ρ_0, as shown in Fig. 5(b) and (c).

2.2 Undercoupled Case (Output Losses Neglected)

The undercoupled case [Fig. 4(b)] is the one in which no half-wavelength shift in the standing-wave-minimum position takes place as frequency passes through resonance, and at resonance the minimum position is shifting away from the microwave tube as frequency is increased.

To find the value of Δd, the following relation is used, as shown in Fig. 5(a):

$$\frac{\Delta d}{\lambda_g} = \pm\left[\frac{1}{4\pi}\tan^{-1}\left(\frac{\rho_0+1}{\rho_0-1}\right)+\frac{3}{16}\right].$$

In Fig. 4(b) points B and C, which are spaced Δd from point O, are projected onto the curve at oblique angles to points A and D. The slopes of AB and CD are such that the effect of frequency on line length is taken into account, as described for Case 1. These slopes are approximately equal to that of the asymptote EF, which the curve approaches at frequencies far from resonance. From points A and D, frequencies f_1 and f_2 are determined. The Q's are found from the following expressions:

$$Q_L = \frac{f_0}{f_1 - f_2},$$

$$Q_U = Q_L\left(1+\frac{1}{\rho_0}\right),$$

$$Q_E = Q_U\,\rho_0 = (1+\rho_0)Q_L.$$

Circuit efficiency is given by

$$\eta_c = \frac{Q_L}{Q_E} = \frac{1}{1+\rho_0}.$$

Cold pulling figure is related to Q_E in the same manner as for the overcoupled case.

If, instead of a plot of the position of the standing-wave-minimum, the curve of voltage SWR as a function of frequency is used, the procedure for finding Q_L is exactly the same as described for the overcoupled

case. However, Q_U, Q_E, and η_c are still related to Q_L and ρ_0 in the manner described above for undercoupled case.

3. Phase of Frequency-Sink Measurements

The phase of frequency sink[7] with respect to an arbitrary reference plane, is the distance, expressed as a fraction of a guide wavelength, of the frequency sink from the reference plane of interest.

A signal generator suitably protected against frequency pulling is connected to a transmission line including a frequency-measuring device and a standing-wave detector. The transmission line is terminated by the microwave tube under test. (See Fig. 1).

The standing-wave detector is used to determine the distance of the nearest minimum in the standing-wave pattern to the reference plane of the tube with respect to which it is desired to determine the phase of frequency sink. This position is expressed in terms of the guide wavelength λ_g as a function of frequency, where measurements are made at frequencies below and above the resonance frequency of the tube under test. Plots similar to those shown in Fig. 7 (overcoupled case) and Fig. 8 (undercoupled case) are obtained. (It is noted that measurements are not performed in the immediate vicinity of tube resonance frequency because of a rapid change of phase in this region and consequent difficulties in obtaining accurate measurements.) A sufficient number of points should be taken to insure that they are falling on the linear portion of the phase characteristic. The phase of sink, is the phase, at the point of resonance frequency, of the straight line drawn through the points of measurement.

In the case of a tunable tube, the measurement may be simplified as follows:

The signal generator is tuned to the frequency f_0 at which it is desired to determine the phase of frequency sink. The tube is tuned to a frequency far away from f_0. The position of the nearest minimum in the standing-wave pattern, d, as explained above, defines the phase of frequency sink f_0.

The procedure is the same in both the undercoupled and overcoupled cases.

4. Measurement of Dispersion Characteristics, Uniform Interaction Circuit

The dispersion characteristic of an interaction circuit is the relationship between the interaction-circuit phase velocity[7] of the various space modes and the frequency of the RF signal propagated by this circuit. The following methods apply to the measurement of circuit dispersion characteristics for the fundamental space mode.

An RF signal source, protected against frequency pulling, is coupled through a suitable transducer to the circuit whose dispersion characteristic is to be determined, as shown in Fig. 9. A signal probe and wavemeter are included, as shown, and the circuit is terminated in a matched load.

The incident wave is reflected by the introduction of a device that can be moved along the axis of the interaction circuit, and whose position is accurately indicated. By moving the reflecting device, the distance d between its positions corresponding to successive minimum readings of the signal-probe indicator is obtained. Then the wavelength along the axis of the circuit is $\lambda_g = 2d$, and the corresponding phase velocity is $v_p = f\lambda_g$, where f is the signal frequency. A plot of v_p vs f can be obtained by changing the frequency of the applied signal.

A method of introducing a reflection is to submerge the circuit in a liquid whose level is accurately monitored. Also a physical short circuit consisting of a metallic cylinder or spring contact may be used.

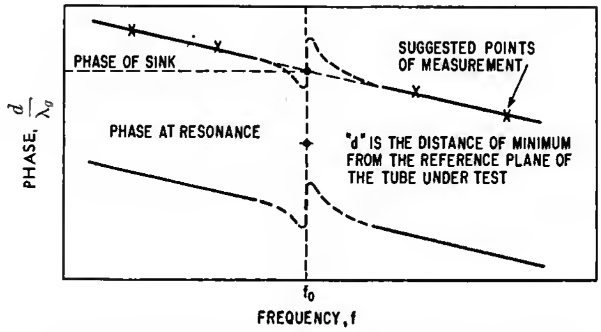


Fig. 7—Phase of standing-wave minimum as a function of frequency (overcoupled case).

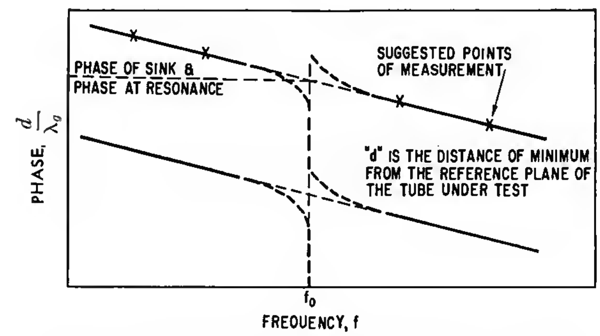


Fig. 8— Phase of standing-wave minimum as a function of frequency (undercoupled case).

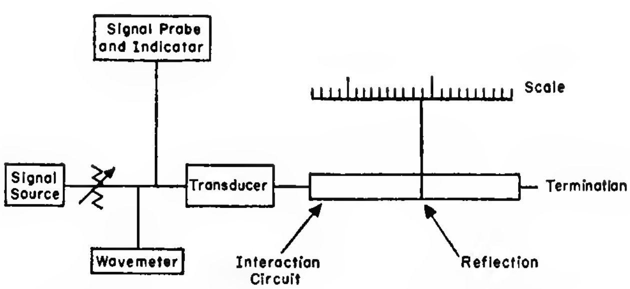


Fig. 9—Circuit arrangement for measurement of dispersion characteristics.

[7] Defined in "IRE standards on electron tubes: definitions of terms, 1957, 57 IRE 7.S2" Proc. IRE, vol. 45, pp. 983–1010; July, 1957.

Variations of those methods have been described in the literature.[8]

In the case of helical interaction circuits, a convenient method of obtaining reflection may be to use a mismatched coupling helix.[9] In some cases, this arrangement has the advantage of permitting measurement of the properties of a helix in an evacuated tube.

B. Microwave Oscillators

5. Power Output

At microwave frequencies power is a fundamental quantity, since current and voltage in a circuit cannot be uniquely defined. The measurement of power is accomplished by conversion of microwave energy to heat.[10–12] The measuring devices are, in general, those which respond to average power, but special techniques are available for the measurements of peak power (see Section 5.2). Since the various power-measuring methods are, in general, applicable to different power levels, the power to be measured can be brought within the range of the particular method by the use of devices of known gain (*e.g.*, calibrated receiver, microwave amplifier) or known loss (*e.g.*, directional coupler.[13] power sampler,[14] calibrated attenuator).

5.1 Measurement of Average Power

Two methods are commonly employed for the measurement of average power.

5.1.1 Continuous-Flow Calorimeter Methods:[15] (Generally used at power levels greater than 1 watt.) In the continuous-flow calorimeter the energy transmitted is absorbed in a continuously flowing stream of fluid. The power in watts P is

$$P = 4.19km\Delta T,$$

where k is the specific heat of the fluid in cal/gm-°C, m is the flow rate of fluid in grams per second, and ΔT is the temperature rise of the fluid in degrees C. Instead of reading m and ΔT directly, calibration of the sytem may be accomplished by substitution of LF power. A system so calibrated consists of two temperature-sensing elements in the fluid stream between which are the microwave absorbing section and the LF power-absorbing section. The RF power is determined by measuring the value of LF power required to produce the same temperature rise as that produced by the RF power. It is convenient to sense ΔT by means of thermocouples or thermopiles.

Consideration should be given to the design of the calorimeter system to insure that the transfer of heat to or from the fluid by extraneous factors is kept negligible, and that the response time of the system is suitable for the particular application. In the use of the substitution method, one source of error due to extraneous heat transfer may be minimized by making measurements with fluid flow in opposite directions and averaging the two.

Because this type of power-meauring device is usually used to terminate a transmission system, it is important that it introduce negligible RF reflection so that substantially all of the power be absorbed, or that the fraction of the power reaching the calorimeter be known.

5.1.2 Bolometer Methods:[16] (Generally used at power levels between 0.1 and 10 mw.) Small values of power may be measured by dissipating the RF energy in temperature-sensitive elements and observing the change in resistance caused by the rise of temperature. Devices that measure power in this manner are designated by the general term *bolometer*. One such device, the barretter, is a piece of fine wire having a high positive coefficient of resistance. Another, the thermistor, a tiny bead of metallic oxides, has a negative temperature coefficient of resistance.

In the most precise measurements of small power, the bolometer is used as one arm of a balanced bridge. The bridge is balanced with direct or LF alternating current through the bolometer. The microwave power then dissipated in the bolometer is measured by the reduction of dc power required to restore balance in the bridge. An unbalanced bridge is often used where less precision is adequate, but where its simplicty of operation is desirable.[16]

For accurate measurement of power, it is necessary that the temperature-sensitive elements be matched to avoid reflections. Because of their nonlinear characteristics, matching under actual operating conditions may be required, especially for pulse measurements.

5.2 Measurements of Peak Power

For measurement of peak power of pulse-modulated microwave sources, two general methods are available.

5.2.1 Pulse-Duty-Factor Method: Measurement of average power by one of the preceding methods can be made and peak power can be calculated by dividing the average power by the pulse-duty factor.[17]

With nonrectangular pulses, peak-pulse power may be somewhat ambigous. The method of measurement is determined by the way in which peak power is defined.[18]

[8] B. Epsztein and G. Mourier, "Definition, mésure et caractères des vitesses de phase dans les systémes a structure periodique," *Ann. Radioélect.*, vol. 10, pp. 64–73; January, 1955.

[9] J. S. Cook, R. Kompfner, and C. F. Quate, "Coupled helices," *Bell Sys. Tech. J.*, vol. 35, pp. 127–128; January, 1956.

[10] C. G. Montgomery, "Technique of Microwave Measurements," McGraw-Hill Book Co., Inc., New York, N. Y., M.I.T. Rad. Lab. Ser., vol. 11, ch. 3; 1947.

[11] M. Wind and H. Rapaport, "Handbook of Microwave Measurements," Polytechnic Inst. of Brooklyn, Brooklyn, N. Y., sect. 4; 1954.

[12] E. L. Ginzton, "Microwave Measurements," McGraw-Hill Book Co., Inc., New York, N. Y.; 1957.

[13] Montgomery, *op. cit.*, ch. 14.

[14] Montgomery, *op. cit.*, ch. 13.

[15] Montgomery, *op. cit.*, pp. 194–220.

[16] Montgomery, *op. cit.*, pp. 84–194.

[17] "IRE standards on pulses: definitions of terms—part 1, 1951," 51 IRE 20. S1, Proc. IRE, vol. 39, pp. 624–626; June, 1951.

[18] G. N. Glasoe and J. V. Lebacqz, "Pulse Generators," McGraw-Hill Book Co., Inc., New York, N. Y., M.I.T. Rad. Lab. Ser., vol. 5., pp. 716–720; 1948.

A typical procedure is to define an equivalent rectangular pulse with minimum departure area.

In the measurement of pulsed power with a bolometer bridge, it is important that the thermal time constant of the bolometer element be much longer than the time interval between pulses so that the change in resistance of the bolometer will accurately indicate the average power. For this reason, a thermistor rather than a barretter is commonly used.

When self-balancing bridges are used, it is important to make sure that the pulse-repetition frequency does not adversely affect the automatic balancing.

5.2.2 The Notch Method: The notch method provides a direct means for measuring instantaneous power. In this measurement a portion of the instantaneous power is detected and the envelope of the RF pulse displayed on an oscilloscope. The power is measured by comparison with a CW signal of the same frequency supplied to the same detector. The CW signal is interrupted when the pulse to be measured occurs. A block diagram of such a system is shown in Fig. 10. The barretter mount and average-power bridge measure half of the power. Accuracy of the measurement is limited by the directional-coupler calibration and by matching of the hybrid junction and terminating elements, as well as by the accuracy of the bridge itself.

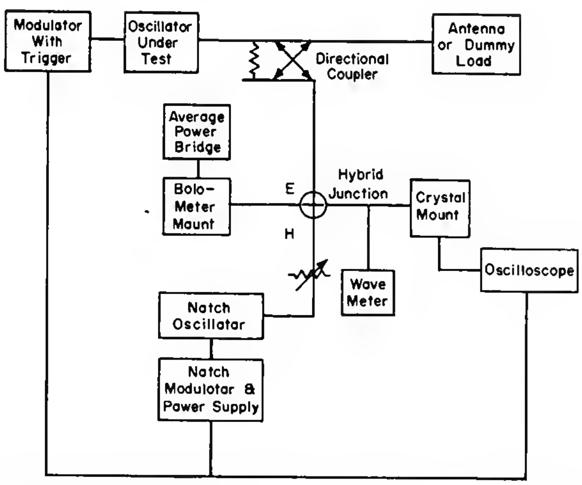


Fig. 10—Block diagram showing the arrangement used in the notch method for measurement of instantaneous power.

6. Methods of Measurement of Frequency

Most microwave-frequency measurements are made by one of two methods, depending on the accuracy required. Further details of these methods may be found in the literature.[19]

6.1 Measurements with an Accuracy of One Part in 10^6 or Better

Harmonics of LF standards are employed. A primary-frequency standard or a secondary-frequency standard, such as a temperature-stabilized crystal oscillator calibrated against a frequency standard, is used. Multiplication is obtained by suitable tube or crystal multipliers, and the known and unknown frequencies are compared by a suitable receiver. This system provides only check points which are multiples of the LF standard.

Additional flexibility may be obtained by using a supplementary frequency oscillator for interpolation between harmonics of the standard oscillator. The frequency of the interpolation oscillator may be measured by a cycle counter. Alternatively, a lower-frequency modulation may be applied to some stage of the multiplier.

6.2 Measurements with an Accuracy of the Order of One Part in 10^3

The most common method of determining the frequency of a microwave signal is by comparing it to the resonance frequency of a calibrated wave meter. The wavemeter is a high-Q tunable circuit that may be either a reaction or transmission type, depending upon the indication of resonance utilized. The two types are shown schematically in Fig. 11.

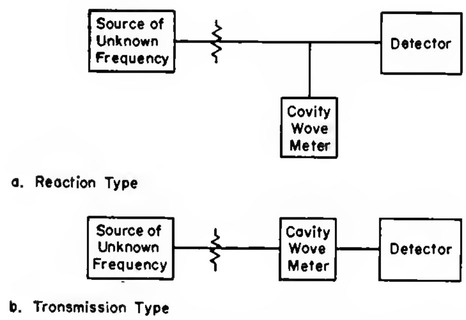


Fig. 11—Arrangements for measuring frequency using calibrated wavemeters.

In the reaction type the wavemeter is in shunt with the detector. When the wavemeter is tuned to the frequency of the signal, a decrease in the detector indication will be noted. In the transmission type, the wavemeter is in series with the detector, and a maximum is noted on the detector when the wavemeter is tuned to the frequency of the signal. The reaction type is more often used because it allows an indication of power present in the circuit even when the wavemeter is not tuned to the correct frequency.

The wavemeter should be loosely or directionally coupled to the system so as not to pull the frequency of the oscillator because of its reactance; also its loaded Q should be high if considerable precision is desired in adjusting the resonance frequency exactly to the frequency of the signal. A loosely or directionally coupled wavemeter will produce a smaller dip if it is a reaction type, or transmit a smaller percentage of the power if it is a transmission type, and the resonance point may be

[19] M. Wind and H. Rapaport, "Handbook of Microwave Measurements," Polytechnic Inst. of Brooklyn, Brooklyn, N. Y.; 1954.

difficult to observe if a wide range of frequency must be searched. Thus, some compromise between precision and ease of use must be made. If very precise measurements are to be made, the dependence of the resonance frequency on temperature and humidity should be known or the device should be designed to minimize these effects as by choice of materials and hermetic sealing.

7. Method of Measurement of Microwave Local-Oscillator Noise

The purpose of this measurement is to obtain a measure of the component of local-oscillator noise that will contribute to the noise of a receiver having a specified intermediate frequency and IF bandwidth.

The noise-to-signal ratio of a local oscillator P_n/P_s was chosen in this discussion as the most useful noise parameter because in practice one will know how much local-oscillator power is directed to the mixer crystal in a receiver installation. By multiplying the oscillator power by the noise-to-signal ratio, one calculates the noise power directly. The noise-to-signal ratio is not a function of oscillator performance alone, but depends also on intermediate frequency and the bandwidth of the receiver with which the local oscillator is used.

In the following discussion where the device is considered only as a local oscillator, two assumptions are made:

1) The power output of the local oscillator is very large compared with that of any signal with which it might beat, including its own side frequencies; that is, the noise-to-carrier ratio of the local oscillator is very small, and the ratio of local-oscillator power to the power of any signal received is very large.
2) The bandwidth of the IF amplifier is much less that the intermediate frequency itself.

The noise power P_n is contributed by the two bands separated from the center of the oscillator spectrum by the intermediate frequency of the receiver.[20] (See Fig. 12.) $S(f)$ is the power in watts per cycle per second of the oscillator. The two noise bands which comprise P_n are shown shaded in the diagram. The measurement of local-oscillator noise may be made by comparing it with the noise in the receiver. Here the receiver consists of a mixer, an intermediate-frequency amplifier, detector and power-output indicator. To measure the noise of the receiver plus the local oscillator, the system shown in block diagram form in Fig. 13 is used. First, with the noise source inoperative, a reading P_1 is registered on the power indicator. Second, with the noise source functioning, a second reading P_2 is taken. If the power from the noise source is $bkT\Delta f$, the receiver noise figure F, is

$$F = \frac{b-1}{a\left(\dfrac{P_2-1}{P_1}\right)}.$$

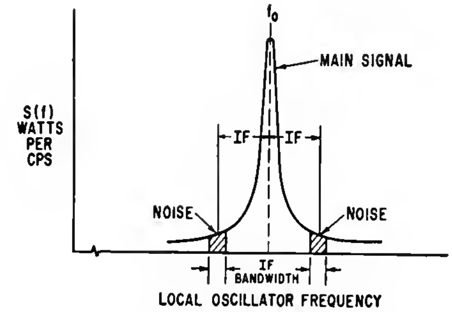


Fig. 12—Local-oscillator spectrum.

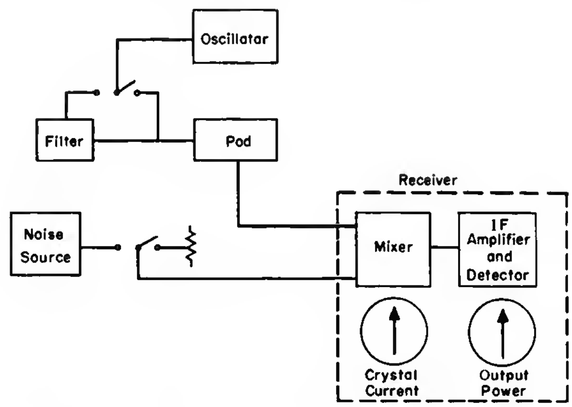


Fig. 13—System for measurement of local-oscillator noise.

The quantity a is the loss between the noise source and the mixer expressed as a numerical ratio $a \geq 1$; k is Boltzmann's constant; T is temperature in degrees Kelvin, Δf is the receiver IF bandwidth. Thus b is the factor by which the noise power of the noise source exceeds that of a matched termination at the reference temperature.

To measure the receiver noise figure alone, we eliminate the noise contributions of the local oscillator by means of a narrow-band filter in the local-oscillator circuit that passes the main local-oscillator power but greatly attenuates the noise sidebands. The filter can be a high-Q transmission-type cavity filter or waveguide filter tuned to the local-oscillator frequency. The frequency of the oscillator should be sufficiently stable so that it remains within the bandwidth of the filter.

The noise-figure measurement is carried out twice: first, using the local oscillator to be tested to determine F, and second, using the local oscillator and filter combination, which can be regarded as a *noise-free* local oscillator, to determine F_0. If

$$a = \frac{F}{F_0},$$

[20] D. Hamilton, J. Knipp, and J. Kuper, "Klystrons and Microwave Triodes," McGraw-Hill Book Co., Inc., New York, N. Y., ch. 17; 1948.

the noise-to-carrier ratio of the local oscillator tested is

$$\frac{P_n}{P_s} = (a-1)\frac{kT\Delta f}{P_x}F_0,$$

where P_n/P_s is the noise-to-carrier ratio and P_x is the local-oscillator power reaching the mixer. In each measurement it is essential that P_x be unchanged, as indicated by the average crystal current. It is also essential that P_x be measured accurately and that it be continuously monitored by the mixer current indicator.

When the receiver is still relatively noisy as compared with the unfiltered oscillator, that is, when $(a-1)$ is very small, it may be desirable to add a filter which will attenuate the carrier as compared with the sidebands.[21–23] If the ratio of the filter attenuation of f_0 to that at $f_0 \pm f_i$ is K, the noise-to-carrier ratio is

$$\frac{P_n}{P_s} = \frac{1}{K}(a-1)\frac{kT\Delta f}{P_x}F_0.$$

8. Transmitting-Oscillator Noise

The measurement of noise in a microwave-oscillator tube differs considerably from amplifier measurements. The output of most oscillators can be represented by one or more of the following models:

1) A noiseless carrier in the presence of a noise background.
2) A carrier, amplitude-modulated by noise.
3) A carrier, angle-modulated (interpreted as either phase or frequency modulation) by noise.

In general, all of these are present with varying amounts of correlation. In practice, a given system is usually affected by only one, with the other negligible. The measurement that is most meaningful depends on the characteristics of the system in which the tube is to be used.

Factors that must be considered in interpreting the results of the noise measurements include power-supply contributions to noise, detector and amplifier noise, stray fields, and microphonics. These effects are often much larger than electronic noise and are very difficult to eliminate.

The measurements to be described apply to transmitting oscillators sufficiently noisy so that the detector noise does not interfere with the measurements. Usually this corresponds to the case where the carrier power exceeds the noise power by less than 80 db.

8.1 Spectrum Measurement

Measurement of the power spectrum is accomplished by means of the system shown in Fig. 14. A very small amount of power from the oscillator to be tested is mixed with a large amount of local-oscillator power to

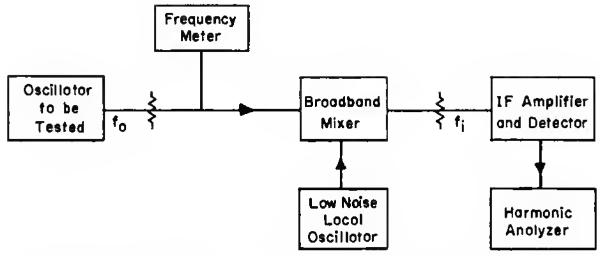


Fig. 14—Arrangement for measurement of power spectrum or of angle-modulation noise.

reproduce the power spectrum of the oscillator under test at a much lower frequency, the intermediate frequency f_i.

The spectrum shape is detected by an AM receiver having a bandwidth narrow compared with the spectrum being measured. If the receiver is tunable over the spectrum width to be measured, the spectrum shape can be determined by observing power output of the receiver as a function of receiver frequency. The spectrum would be a plot of power density in watts per unit bandwidth as a function of frequency. A carrier-to-noise ratio is obtained by integrating the carrier portion and the noise portion of the spectrum over the frequency bands specified for each, and expressing as a result the quotient of the two.

The same result may be achieved by keeping the frequency of the AM receiver fixed and tuning the local oscillator in small increments, recording AM receiver power output as a function of local-oscillator frequency. In either case, the noise output of the local oscillator and the mixer must be very small compared with that of the tube under test. The power output of the local oscillator must also be very large compared with the power at the mixer of the oscillator under test.

Resolution in frequency may be limited either by the stability of the local oscillator or by the bandwidth of the receiver, and is usually limited by the local oscillator. It is possible either to observe the varying output of the receiver as a function of frequency or to keep the receiver output constant by means of a calibrated attenuator, using attenuator settings as the measure of power output at each frequency setting.

8.2 Amplitude-Modulation Noise

Amplitude-modulation noise may be measured by the system in Fig. 15. A power sample is applied to a calibrated detector, the direct current from the detector indicating the average power of the oscillator. Alternating components that are caused by amplitude modulation of the detector current are amplified in an ac amplifier. The alternating current measured at the calibrated amplifier output can be referred to the input and compared with the detector current. Since the dc detector output is related to the average power of the oscillator under test, the ac power at the amplifier input can be converted into absolute units of AM noise power over

[21] G. C. Dalman and D. Ortiz, "Measurements of microwave local-oscillator noise," *Proc. NEC,* vol. 9, pp. 833–840; 1953.
[22] D. C. King, "Measurements at Centimeter Wavelength," D. Van Nostrand Co., Inc., New York, N. Y., ch. 3; 1952.
[23] See Sections 7, 8, and 22.

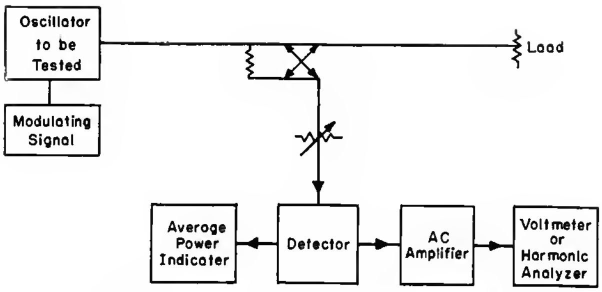


Fig. 15—Arrangement for measurement of AM noise.

the bandwidth of the ac amplifier. The detector must be linear or, for large carrier-to-noise ratio, the incremental calibration of the detector must be known at the dc operating level of the detector. Calibration of the amplifier for the particular operating level minimizes error.

8.3 Angle-Modulation Noise

Angle modulation (phase or frequency modulation) can be measured by means of the system of Fig. 14, where the IF amplifier and detector in the diagram are replaced by an FM receiver in which limiting is used to suppress AM effects in the measurement. The AM-FM conversion characteristic of the limiter must be such that no additional frequency modulation is introduced. FM noise output may be determined from the output of such a receiver when the discriminator characteristics are known. Such a measurement requires that the local oscillator be much more stable in terms of angle modulation than the oscillator under test, a requirement which is often difficult to meet.

To eliminate the effects of the local oscillator, a system of direct detection, shown in Fig. 16 can be used. The oscillator to be tested is operated into a suitably terminated transmission line. Two power samples are taken from the line, one to be used in an AM-FM channel, the other to be used in an AM channel. The AM-FM channel consists of a variable attenuator, a discriminator, and a detector. The output voltage of the detector is mainly direct current with a small alternating current superimposed on it. The alternating current is made up of two components, one proportional to amplitude variations and one proportional to frequency deviations. The output of the AM channel is also a direct voltage with a small alternating voltage superimposed on it, the alternating voltage being proportional to amplitude variations.

The amplitude-variation signals in each channel are balanced out in a mixing transformer. The direct voltages at each detector are made equal, and the detector outputs are connected to the transformer primary in such a way that AM components in each channel cancel. The output of the ac amplifier is then a voltage which is proportional to frequency deviation. This system can measure AM and FM noise separately, but is accurate only when there is little or no correlation between AM and FM. Where AM noise is to be measured, the AM channel alone can be used. (Means for checking AM-FM correlation are included in Gottschalk.[24]) Where AM noise is very small, the cancelling process in the transformer can be omitted, and the AM-FM channel can be used to measure FM noise. When appreciable AM-FM correlation exists, the sensitivity of the discriminator must be increased so that its output is predominantly due to FM noise.

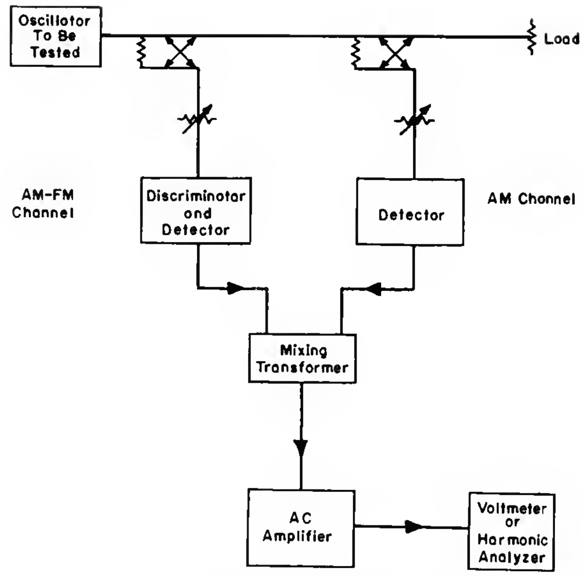


Fig. 16—Arrangement for angle-modulation noise measurement—a system of direct detection eliminating effects of the local oscillator.

The frequency deviations can be measured quantitatively by calibrating the test apparatus with the forced frequency modulation of a known oscillator. One can measure the rms frequency deviation over the video bandwidth by measuring the output of the ac amplifier with an rms voltmeter and making the proper conversion of frequency. One can also measure the spectrum of the FM noise by means of a harmonic analyzer at the ac amplifier output.[25,26]

9. Loading Effects

Performance of an oscillator with respect to variations in load impedance is presented by means of a load-impedance diagram.[27] The most usual form of load-impedance diagram is the Rieke diagram,[28] which shows "contours of constant power output and constant frequency drawn on a polar diagram whose coordinates represent the components of the complex reflection coefficient of the oscillator load impedance." It is useful to

[24] W. M. Gottschalk, "Direct detection measurements of noise in CW magnetrons," IRE Trans. on Electron Devices, vol. ED-1, pp. 91–98; December, 1954.
[25] R. A. La Plante, "Development of a low-noise X-band CW klystron power oscillator," IRE Trans. on Electron Devices, vol. ED-1, pp. 99–103; December, 1954.
[26] R. V. Pound, "Frequency stabilization of microwave oscillators," Proc. IRE, vol. 35, pp. 1405–1415; December, 1957.
[27] Defined in "IRE standards on electron tubes: definition of terms, 1957," 57 IRE 7.S2, Proc. IRE, vol. 45, p. 999; July, 1957.
[28] *Ibid.*, p. 1004.

show regions of instability,[29] poor spectrum in modulated tubes, and other forms of unsatisfactory operation. In general, the Rieke diagram is presented for one set of input conditions, but may be useful as an indication of what will happen for other input conditions. In general, the anode current in magnetrons and the anode voltage in klystrons is kept constant. To minimize errors due to thermal effects, the measurements are made so that resonator temperatures are held as constant as possible.

A suitable arrangement for the measurement of power and frequency as a function of loading is shown in Fig. 17.

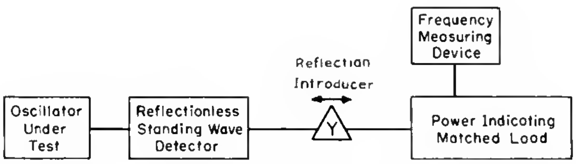


Fig. 17—Arrangement for measurement of power and frequency.

The reflection introducer is a device with which the phase of an introduced standing wave can be varied independent of the magnitude of the standing wave, and, if possible, vice versa. In order to avoid errors in power measurement, it is necessary that the power loss introduced by the reflection introducer be negligible.

The standing-wave detector, which is used to measure the phase and magnitude of the reflection from the reflection introducer, may be omitted if the reflection introducer has been previously calibrated. It is usually convenient to refer the phase of reflection to some designated part of the tube.

The procedure for making a Rieke diagram is as follows:

1) Make power and frequency measurements under conditions of no reflection ("matched load").
2) With a small magnitude of reflection introduced, measure power output and frequency at points corresponding to a number of phases of reflection.
3) Repeat these measurements for successively higher magnitudes of reflection, up to as high magnitudes as may be of interest.
4) Make a polar plot of magnitude and phase of reflection, indicating lines of constant frequency and constant power output.

The *pulling figure* is measured by setting the reflection introducer for a reflection coefficient of 0.20 and observing the frequency extremes reached as the phase angle of the reflection coefficient of the load impedance varies through 360°.

The *frequency sink* of an oscillator is the locus on the Rieke diagram of points of maximum rate-of-change-of-frequency with phase for all reflections. Maximum power output usually occurs near the frequency sink. Unstable operation is often encountered near the frequency sink, especially with high values of load reflection.[30–33]

10. Methods of Measurement of Mechanical Tuning Characteristics of Microwave-Oscillator Tubes

The measurements usually fall into six general categories: calibration, resettability, stability, life, starting force, and operating force. Specific measurements in each of these categories are:

1) Calibration
   a) Sensitivity
   b) Range
2) Resetability
   a) Backlash
   b) Hysteresis
   c) Creep
3) Stability
   a) Shock
   b) Vibration
   c) Temperature change
   d) Ambient pressure change
4) Life
5) Starting force (or torque)
6) Operating force (or torque).

These items are inter-related. The order of listing is significant only in grouping of the characteristics that are tested in similar manner. It is assumed that motion of the drive mechanism will be measured from a prescribed point with any suitable motion-measuring device.

Fig. 18 is a block diagram of the basic circuit arrangement common to the required measurements. Components should be arranged to provide impedance matching suitable for the measurement accuracy desired. Frequency pulling, for example, caused by wavemeter resonance should be reduced to a sufficiently small quantity by loose coupling.

[29] The performance of an oscillator when connected to a load that is frequency-sensitive may be marked by unstable operation at some frequencies. This effect, when encountered with a load consisting of a long length of transmission line terminated in a mismatched load, is termed *long-lines effect*. The critical *length of line* for an oscillator for a specified reflection at the termination is defined as the minimum length of line that will cause the operating frequency to become a discontinuous function of load phase. This length decreases with increasing oscillator pulling figure and with increasing load reflection. Consequently, as short a line as possible should be employed in making measurements of loading effects. See Collins, *op. cit.*, ch. 7, sect. 6; and Montgomery, *op. cit.*, ch. 2, sect. 3.

[30] G. B. Collins, "Measurements of the resonant systems," in "Microwave Magnetrons," McGraw-Hill Book Co., Inc., New York, N. Y., M.I.T. Rad. Lab. Ser., vol. 6; 1948.

[31] H. J. Reich, P. F. Ordung, H. L. Krauss and J. G. Skalnik, "Microwave Theory and Techniques," D. Van Nostrand Co., Inc., New York, N. Y., ch. 10; 1953.

[32] J. C. Slater, "Microwave Electronics," D. Van Nostrand Co., Inc., New York, N. Y.; 1950.

[33] C. G. Montgomery, "Technique of Microwave Measurements," McGraw-Hill Book Co., Inc., New York, N. Y., M.I.T. Rad. Lab. Ser., vol. 11; 1947.

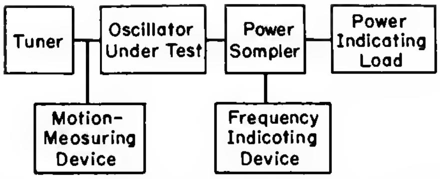


Fig. 18—Basic arrangement for measurement of mechanical tuning.

10.1 Calibration

Fig 18 shows an arrangement that will provide data for a curve of tuner motion vs frequency. This curve is the *calibration* and the slope of the curve is the *sensitivity* at any given point. It is generally necessary to make the entire calibration curve without reversing the direction of tuning, and to specify in which direction it was made. This is done to eliminate the effects of hysteresis and backlash. The *mechanical tuning range* of an oscillator is the frequency interval of continuous tuning over which the essential characteristics remain within specified limits.

10.2 Resetability

This is the ability of the tuner to produce the same operating frequency for the same tuner setting. Backlash is the amount of motion of the tuner mechanism that produces no frequency change on reversal of tuner motion direction. It may be measured by reversing the direction of tuning and observing the amount of motion that takes place before any change of frequency is observed. *Hysteresis* describes any effects observed when a given tuner setting is approached from opposite directions. It may include *backlash*. *Creep* is the variation of the tuner calibration curve with repeated cycling of the tuner. The observed change in calibration after a specified number of tuning cycles is a measure of *creep*. A tuning cycle is a complete excursion and return of the tuner through a specified range.

10.3 Stability

Mechanical tuning stability with regard to shock, vibration, temperature change, and ambient pressure change is measured with the arrangement shown in Fig. 18. The tube is subjected to the conditions required, and the resultant frequency change is noted.

10.4 Life

With normal voltages applied, the tuner is moved through its cycle continuously. Periodic checks are made of the characteristic under study. The number of cycles before the characteristic departs from the prescribed limits is a measure of life.

10.5 Starting Force (or Torque)

Here the motion-measurement device of Fig. 18 is supplemented by a device that measures force. The force required to produce an initial frequency shift is the *starting force*.

10.6 Operating Force (or Torque)

With an arrangement as above, the force required to maintain a specified rate of change of tuning is measured.

11. Electrical Tuning

Ordinarily an electrically tuned oscillator is tuned by altering one electrical input. Special cases may arise where tuning takes place by altering two or more electrical inputs together according to a prescribed schedule. Listed below are measurable electrical-tuning characteristics normally of interest. The terms used herein have the same meaning as they have in Section 10.

1) Calibration
   a) Range
   b) Sensitivity
2) Resetability
   a) Hysteresis
   b) Creep
3) Tuner input impedance
4) Response time.

Point-by-point characteristics may differ markedly from dynamic swept-frequency characteristics because of incidental thermal effects in electrically tunable tubes. Fig. 19 is a block diagram of the basic circuit for making the required measurements on a point-by-point basis. Fig. 20 is employed for swept-frequency measurements. Components in each case should be arranged to provide impedance matching suitable for the measurement accuracy desired. For example, loose or directional coupling between the oscillator under test and the wavemeter is necessary to avoid pulling of the oscillator when measuring frequency.

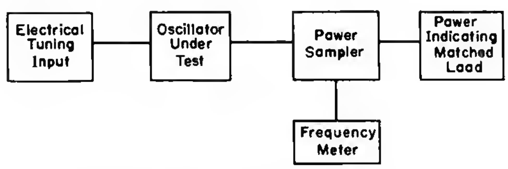


Fig. 19—Block diagram of basic circuit for point-by-point measurement of electrical tuning.

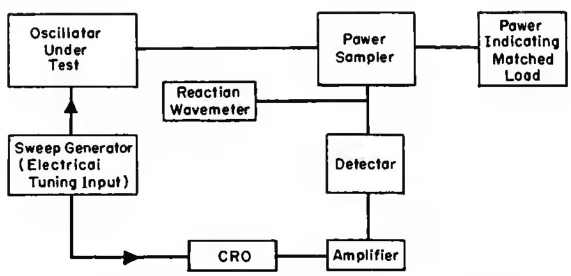


Fig. 20—Block-diagram arrangement for swept-frequency measurements of electrical tuning.

The rate of change of frequency vs the input parameter at any given value of the input parameter is the electrical-tuning sensitivity at that value. To eliminate the effects of hysteresis, it is generally necessary to take the entire calibration curve without reversing direction of tuning and to specify in which direction it was made. To measure effects due to hysteresis, the oscillator should be tuned continuously over its entire range in one direction, and then in the opposite direction.

It is of interest to know the tuner input impedance during operation. In some types of tubes the voltage-current relationship may be described by an equivalent linear passive-circuit element, while in others it may be necessary to indicate specifically the voltage-current characteristics. As compared with the point-by-point characteristic, the swept-frequency characteristic may require the addition of equivalent reactive circuit elements. In most cases the tuner input impedance is measured by conventional LF methods.

The response time may be measured by applying a step change of the input to the tuning electrode and observing directly the change of frequency. The response time may be also be observed by noting the reduction in the frequency deviation as the frequency of a constant amplitude-modulating signal is increased.

12. Modulation of CW Oscillators

By variation of a control parameter, *e.g.*, electrode voltage, magnetic field, tuner position, etc., of a microwave oscillator, the amplitude and/or frequency of the oscillator can be changed. This property can be used to modulate the carrier in a desired manner. Electrical tuning, which is closely related to frequency modulation is described in Section 11.

12.1 Amplitude Modulation

Basically, the AM characteristic shows how the output power varies as a function of one or more control parameters. For a single control parameter, the AM sensitivity is defined as the slope of the power output vs control-parameter characteristic at the value of the control parameter of interest. To measure the AM sensitivity under dynamic conditions, an ac variation is applied to one of the oscillator controls, as in Fig. 21. A power sampler is applied to a detector, amplified by a dc amplifier with sufficient bandwith, and fed into an appropriate indicator. The detector output will contain a dc component proportional to the average power of the oscillator and ac components due to the amplitude modulation. The ac power may be determined by reference to the average power and the AM sensitivity expressed in absolute units if the detector and amplifier are calibrated, or a calibrated attenuator may be employed to measure the power ratio. It should be noted that the modulating frequency, oscillator frequency and load conditions may strongly affect the modulation characteristic.

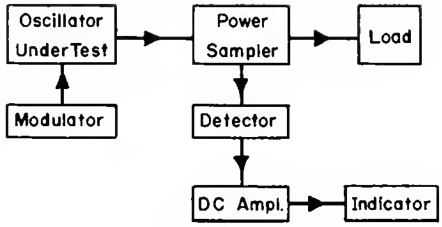


Fig. 21—Equipment for AM sensitivity measurement.

Strictly speaking, modulation sensitivity is defined only over a range for which the tube-modulation characteristic is essentially linear. However, in many cases it may be desirable to measure an effective modulation sensitivity for a nonlinear modulation characteristic. In such cases, this effective sensitivity will be a function of the amplitude of the modulating signal. Under these conditions it may also be desirable to measure distortion, as described in Section 12.3.

12.2 Frequency Modulation

The FM characteristic describes the frequency change as a function of one or more control parameters. In the case of a single control parameter, the FM sensitivity is defined as the slope of the frequency vs control-parameter characteristic at the value of control parameter of interest. To measure the FM sensitivity under swept-frequency conditions, an ac variation is applied to one of the oscillator controls. Three general methods are applicable for the measurement of FM characteristics.

12.2.1 Carrier-Null Method: A sample of the oscillator output is fed into a spectrum analyzer, and the first minimum of the carrier is obtained by varying the amplitude and/or frequency of the applied modulation. The value of the modulating control parameter is gradually increased from zero until the amplitude of the carrier as seen on the spectrum analyzer goes through a minimum. The FM sensitivity can be calculated under these conditions from the relation[34]

Modulation sensitivity

$$= \frac{2.405 \times \text{Modulating frequency}}{\text{Peak amplitude of modulating parameter}},$$

where the peak amplitude of the modulating parameter is the value obtained from the first minimum of the carrier. This method is useful for measuring deviations small enough to assure linearity and small incidental amplitude modulation. (See Fig. 22.)

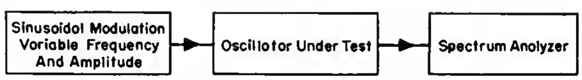


Fig. 22—Arrangement for measurement of FM characteristics (carrier-null method).

[34] "Reference Data for Radio Engineers," IT&T Corp., New York, N. Y., 4th ed. p, 535; 1956,

12.2.2 Discriminator Method: For narrow-band evaluation, output of the oscillator is fed to a calibrated discriminator of either the microwave or IF type. In either case, some means of suppressing the effects of incidental amplitude modulation is required, either by limiting[35] (Fig. 23), or by a compensating microwave detector[36,37] (Fig. 24). The frequency deviations which may be measured by this technique are limited by the bandwidth of the discriminator.

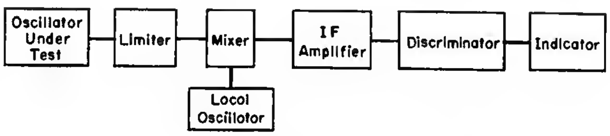


Fig. 23—Arrangement for measurement of FM characteristics (discriminator method).

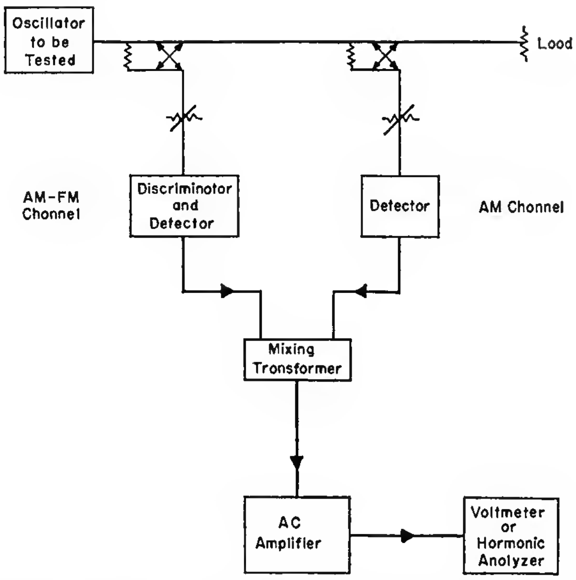


Fig. 24—Equipment for measurement of FM characteristics (discriminator method) using compensating microwave detector.

12.2.3 Frequency-Meter Technique: This is equivalent to the measurement of electrical tuning under dynamic conditions.[38]

The comments on the effect of linearity, frequency and load under AM apply equally to the FM case.

12.3 Distortion

Distortion is produced when variations in the output amplitude, frequency or phase are not linearly related to variations of control parameter. Distortion can be expressed either in terms of the deviation of modulation characteristic from linearity or in terms of the harmonic content introduced into the output. The harmonic distortion can be determined by applying the modulated signal to an appropriate detector for converting the amplitude, frequency or phase variation into a voltage variation which may be analyzed for harmonic content by means of a wave analyzer. When this method is used, extreme care is necessary in the detector design to insure that additional harmonics are not introduced by the detector.

12.4 Carrier-Frequency Shift

In either amplitude or frequency modulation, a phenomenon known as *carrier-frequency shift* may occur. The amount of shift may be a function of the amplitude and/or frequency of the applied modulation. If the modulating amplitude and/or frequency is swept over the required test range, any deviation of the carrier frequency is readily observed on a spectrum analyzer.

13. Pulsed-Oscillator Measurements

13.1 Measusement of RF Spectrum

The RF output of a pulsed oscillator may be expressed, by the method of Fourier analysis, as the summation of the voltages of sine and cosine waves with incremental frequency differences over all frequencies from zero to infinity. The amplitude of these voltages as a function of frequency is the voltage spectrum. The square of the magnitude of the voltage spectrum is the power spectrum.

For a rectangular pulse of duration τ with no frequency or amplitude modulation during the pulse, the power spectrum is given by

$$P(f) = A \frac{\sin^2 (f_0 - f)\tau\pi}{[(f_0 - f)\tau\pi]^2},$$

where f_0 is the frequency of the oscillation during the pulse.

Examination of this equation, sketched in Fig. 25, reveals that the power spectrum consists of a main lobe centered around $f = f_0$, with minor lobes, which are rapidly attenuated, extending to higher and lower frequencies.

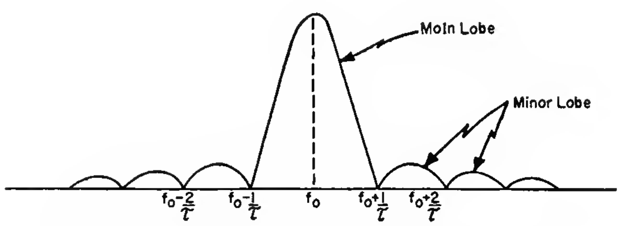


Fig. 25—Power spectrum of ideal rectangular pulse.

Actual pulse shapes employed to modulate RF oscillators deviate from the rectangular form. Since oscillator frequency is usually sensitive to applied voltage, frequency modulation of the RF output is produced. These effects can combine to produce various

[35] For example, see F. B. Frank and G. Wade, "Traveling-wave tube limiters," IRE Trans. on Electron Devices, vol. ED-4, pp. 148–152; April, 1957.
[36] R. A. La Plante, "Development of a low-noise X-band, CW klystron power oscillator," IRE Trans. on Electron Devices, vol. ED-1, pp. 99–103; December, 1954.
[37] R. V. Pound, "Frequency stabilization of microwave oscillators," Proc. IRE, vol. 35, pp. 1405–1415; December, 1947.
[38] See Section 11, Electrical Tuning.

forms of spectrum deformation, such as variation of the height and width of the main lobe and minor lobes and the loss of zero intercepts. The basic characteristics of the spectrum are defined by the RF bandwidth which is the width, in megacycles per second, of the main lobe measured at a standard level below its peak (usually 6-db down), and the minor-lobe ratio, which is the db amplitude ratio of the highest minor lobe relative to the main lobe.

The measurement of spectrum characteristics is usually made by means of a spectrum analyzer, or RF spectrometer.[39] This consists of a highly selective superheterodyne receiver whose output is brought to the vertical deflection plates of the cathode-ray oscilloscope. The receiver is slowly swept in frequency while the horizontal sweep of the cathode-ray tube is varied in synchronism. Each time the oscillator under study is pulse modulated, a transient appears on the oscilloscope whose amplitude is proportional to the value of the power spectrum at the instantaneous frequency to which the receiver is tuned. The envelope of the succession of transients which appear upon the screen is thus proportional to the power spectrum.

For good resolution of spectrum-frequency components, the receiver bandwidth must be narrow compared with $1/\tau$. However, the maximum RF bandwidth that is consistent with resolution requirements should be employed to maximize the allowable pulse-repetition frequency.

13.2 *Pulse Jitter and Missing Pulses*

Pulse jitter of a microwave oscillator arises from the inability of the oscillator to generate pulses of identical characteristics. Three types of pulse jitter usually identified are starting time, amplitude, and frequency jitter, corresponding, respectively, to variations in relative starting time, amplitude, or carrier frequency. These jitter variables may be expressed either in peak-to-peak, average or rms values on the basis of the variation of each individual pulse from a mean value or the variation of a particular pulse from the preceding pulse. Methods similar to those described here may be used to measure pulsewidth jitter and trailing-edge jitter.

After all obvious cyclic variations have been removed, the distribution of the remaining random variations is nearly Gaussian. The distribution function, plotted as cumulative per cent vs the particular jitter variable on normal probability paper, will therefore yield a straight line. For example, Fig. 26 shows the relation between an actual jitter, in this case time, and a distribution plot of cumulative per cent vs variation. The rms value of the jitter variable, defined as the standard deviation[40] σ, is given by the difference in the variable between the 50 and 84.13 per cent points. (If the occurrences are not Gaussian, the true pulse-to-

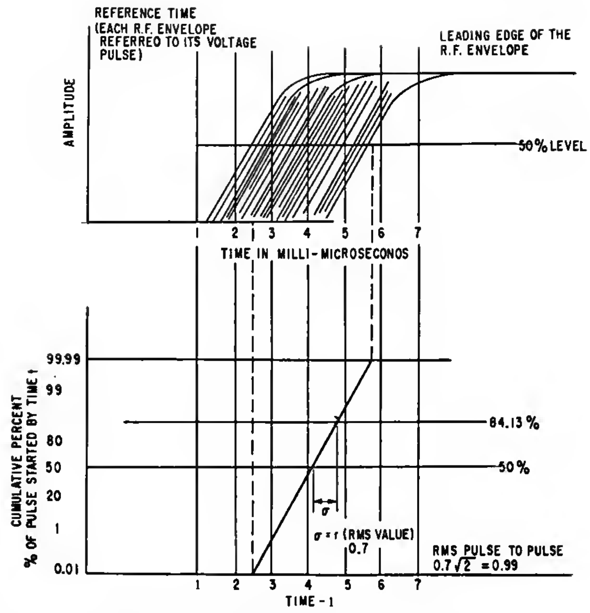


Fig. 26—Distribution plots of amplitude and of cumulative per cent vs time as a jitter variable.

pulse rms value will not be given by the difference mentioned above. An estimate of the true value can be obtained if the amount of cyclic content is known.)

For most measurements the deviation from the mean is easier to measure than the variation between successive pulses. The value measured from the mean is converted to a pulse-to-pulse value[41] by multiplying by $\sqrt{2}$.

Fig. 27 shows one system which may be used to measure time, frequency and amplitude jitter. A sample of the RF energy is fed simultaneously to a microwave interferometer[42] and a precision attenuator and detector. The detector may be run with or without a bias. The output of the interferometer is used to obtain frequency jitter, the output of the unbiased detector alone is used to obtain the time jitter, and the output of the biased detector is used to obtain amplitude jitter. The remainder of the system is common to all measurements.

13.2.1 Time Jitter: Time jitter is the variation of the time incidence of the leading edges of two successive output pulses of the detector. In an actual system this jitter includes the effects of both the transmitter and receiver, as well as of the microwave oscillator. To measure the contribution of the oscillator alone, time jitter is measured as the variations of the interval between the leading edge of the voltage pulse applied to the tube and the leading edge of the resulting RF envelope, generally at the 50 per cent level.

As shown in Fig. 27, the output of the detector is fed into a video amplifier and a precision variable delay line into a special coincidence circuit. Since the time

[39] Montgomery *op. cit.*,[33] ch. 7.
[40] "Reference Data for Radio Engineers," *op. cit.*, p. 983.

[41] The $\sqrt{2}$ results from the addition of two equal deviations.
[42] H. P. Raabe, "Measurement of instantaneous frequency with a microwave interferometer," Proc. IRE, vol. 45, pp. 30–38; January, 1957.

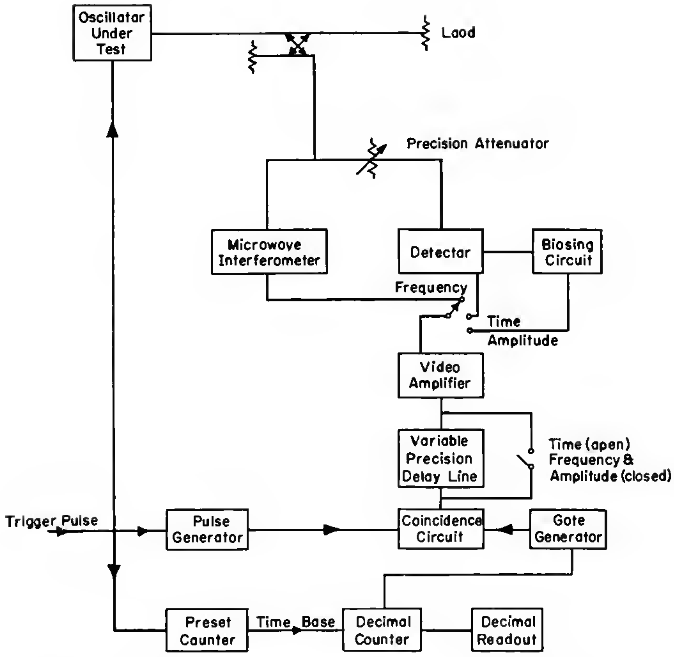


Fig. 27—An arrangement for measuring jitter variables, time, frequency, and amplitude.

jitter is measured with reference to some point of the applied voltage pulse, this voltage pulse is used to trigger a pulse generator. The characteristics of this pulse generator include a variable pulsewidth and low jitter between the triggering pulse and the output pulse. The relationship between the signal output of the pulse generator and the RF envelope is shown in Fig. 28. The trailing edge of the output pulse of the pulse generator is combined with the leading edge of the RF envelope in the coincidence circuit. The output of the coincidence circuit is a triangular-shaped wave in which the amplitude is now proportional to time and in which variations in amplitude are proportional to the time variations of the leading edge of the RF envelope. This signal is then fed into a gate generator. The level required to trigger the gate generator is usually set at a value such that the amplitude of the triangular pulse represents the 50 per cent level of the leading edge of the RF envelope. Each time the triangular wave exceeds this level a pulse is generated in the gate generator and fed into a decimal counter. As the delay is varied, the amplitude of the triangular wave increases or decreases, depending upon the direction in which the delay is varied. It is possible to vary the delay line in small increments in order to allow the triangular waveform to vary in amplitude so that the gate generator produces 0 to 100 per cent of the number of applied voltage pulses.

The output of the gate generator is fed into a decimal counter having a time base established from a preset counter, driven by the voltage pulse applied to the oscillator under test. This allows a uniform number of pulses to be used as a base rate.

The procedure used in obtaining the distribution curve of time jitter is to position the variable delay line

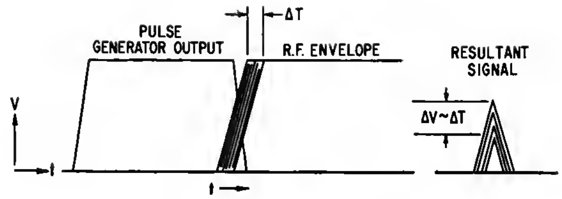


Fig. 28—Time jitter.

so that no counts occur on the decimal counter. The delay is then changed by a small amount. The counter will record the number of pulses exceeding the preset level of the gate generator. The time delay is varied through such a range of values that the count recorded on the decimal counter goes from zero to the maximum or from 0 to 100 per cent of the input pulses. When the increments of time delay are known, it is possible to plot the per cent of pulses occurring on the decimal counter as a function of time. This will yield a plot of cumulative per cent as a function of time jitter (in millimicroseconds generally). From this plot the rms value may be obtained and the pulse-to-pulse rms value computed in the manner discussed earlier.

13.2.2 Frequency Jitter: Frequency jitter is the difference in the oscillator output frequency of two successive pulses. Frequency jitter is measured by means of the arrangement shown in Fig. 27, with the microwave interferometer used as the sensing element. The output of the interferometer is as shown in Fig. 29A; the amplitude of the detected output pulse from the interferometer is a function of frequency. This signal is fed into the video amplifier and then into the coincidence circuit where it is combined with a pulse from the pulse generator (Fig. 29B). The resulting signal is a narrow pulse having amplitude variations proportional to the frequency variations of the output of the microwave device (Fig. 29C). This pulse, which is varying in amplitude, is fed into the gate generator and handled in the same way as for time jitter. However, in the case of frequency jitter the distribution is determined by changing the reference frequency of the interferometer in discrete steps; this results in moving the display shown in Fig. 29C up or down in amplitude. It is then possible to go from 0 to 100 per cent of pulses.

13.2.3 Amplitude Jitter: Amplitude jitter is the difference in oscillator power output of two successive pulses. Amplitude jitter is measured by means of the arrangement shown in Fig. 27 with the biased detector used as the sensing element. The RF envelope is detected and the bottom clipped by the biasing circuit (Fig. 30A). The top of the envelope is then fed through the video amplifier and into the coincidence circuit, where it is combined with the output of the pulse generator shown in Fig. 30B. This results in a very narrow pulse having amplitude variation that is a function of the amplitude variations of the microwave device (Fig. 30C). The signal is then fed into the gate generator and handled in a manner similar to that used for time and frequency

jitter. The actual distribution of amplitude jitter as a function of per cent of pulses is obtained by adjusting a precision attenuator located just before the detector.

Note: It should be pointed out that in time jitter, frequency jitter, and amplitude jitter the distribution as plotted on normal probability paper will result in a straight line only if the occurrences have a Gaussian distribution. A departure from linearity usually indicates a cyclic content.

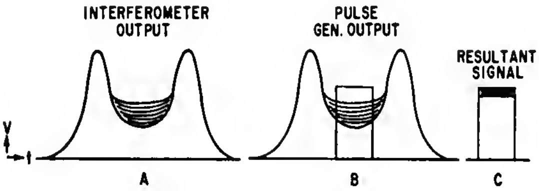


Fig. 29—Frequency jitter.

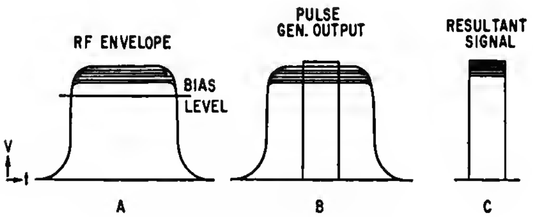


Fig. 30—Amplitude jitter.

13.2.4 Measurement of Number of Missing Pulses: A measure of the stability of a pulsed oscillator is its ability to generate repetitive RF pulses of approximately the same frequency and amplitude. This ability is markedly affected by both the driving source and the load characteristics. Therefore, stability measurements should specify pertinent source and RF load parameters.

The *missing pulse* is defined as one of a repetitive train of RF pulses whose integrated amplitude (energy) is less than some specified percentage (usually 70 per cent) of the normal pulse energy over a specified bandwidth. The missing pulses are counted for a fixed time interval, and the ratio of missing output pulses to total input pulses is expressed in per cent.

The circuit in Fig. 31 can be used to measure the number of missing pulses. A small sample of the RF power is fed to a crystal operated in the region where the output voltage is proportional to input power. The output is fed into a wide-band amplifier and then integrated. The peak amplitude of the integrator output serves as a reference for the following pulses. The time constant of the storage circuit is many times greater than the interpulse time, but small enough so that the reference follows input changes that occur over seconds. In order to select the rejection level, the stored voltage is attenuated by adjustable fixed steps and fed into a comparison circuit that compares this average level with that of an integrated single pulse. If the peak amplitude of the integrated pulse is larger than the average, the output pulse triggers a multivibrator. The multivibrator output is fed to an anticoincidence gate that is also fed by a negative reference pulse. A pulse therefore appears at the output whenever the RF pulse energy is less than a predetermined value. The output is counted by a decade counter. It should be noted that, because of the self bias, this circuit continues to measure the number of missing pulses even when the average RF level undergoes changes caused by RF loading or change in plate voltages.

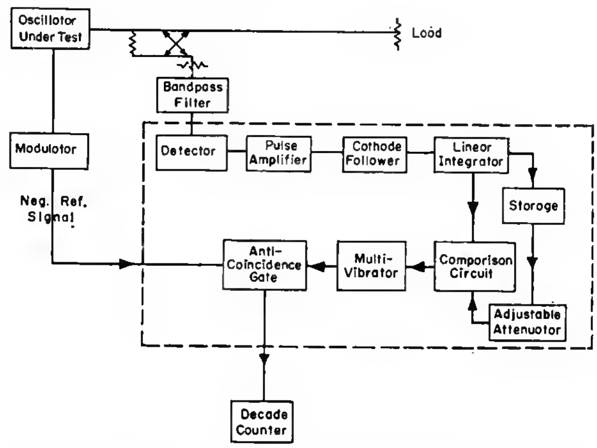


Fig. 31—Arrangement for determination of missing pulses.

Excellent qualitative indications of stability can be obtained by a careful examination of the current pulse,[43] the RF frequency spectrum, or the detected RF envelope.

14. Spurious Oscillations

This section concerns only spurious microwave oscillations generated by microwave oscillators. It excludes oscillations outside the microwave region and X rays.

14.1 CW Oscillators

For CW oscillators, a spurious oscillation is an oscillation that exists in the presence of the carrier at other than the carrier frequency. Such an oscillation is to be distinguished from externally caused modulation components or the output of an unstable tube where a stable carrier is not obtained.

14.1.1 Point-by-Point Measurement: The spurious oscillation output of a CW oscillator can be measured on a point-by-point basis with the oscillator operating at a fixed frequency and the frequency band of interest searched by means of manual adjustment of a tunable selective detector. The measurement system to be employed must of necessity depend upon the relative magnitudes and frequencies of the main oscillation and the spurious oscillations to be studied. One generally suitable approach is the analysis of a sample of the oscilla-

[43] Glasoe and Lebacqz, *op. cit.*, pp. 676*ff*.

tor output by means of a microwave spectrum analyzer.[39] A signal generator may be used as a comparison standard to measure the relative amplitudes of the main oscillation and the spurious oscillation. Care must be taken to identify spurious responses of the analyzer produced by its mixer.

Spurious signal outputs of high-power oscillators may be measured by the use of a transmission wavemeter and microwave power bridge or other form of simple detector without resorting to superheterodyne receivers. The sensitivity and accuracy of such measurements in the vicinity of the carrier frequency are limited by the response of the wavemeter. Spurious outputs may be a function of the loading of the tube at the carrier frequency and also at the frequency of the spurious oscillation.

14.1.2 Swept-Frequency Measurement: Since the characteristics of many electrically tunable oscillators flucutate markedly throughout their tuning range, it is usually desirable to examine their output over a continuous broad frequency band, as the tube is continuously tuned over all or a part of its electrical tuning range. One method of accomplishing this is shown in Fig. 32.

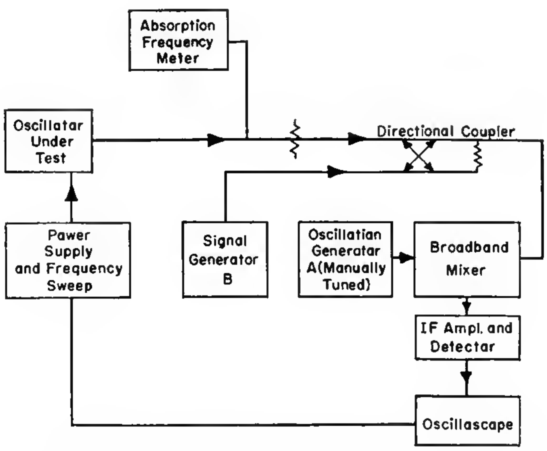


Fig. 32—Equipment for measuring spurious signals.

Oscillation generator A, the mixer, the IF amplifier and the oscilloscope constitute a manually tuned narrow-band receiver whose output as a function of carrier frequency is displayed on the oscilloscope. If the same sweep voltage is employed for the tube under test and for the oscilloscope, every point on the horizontal axis of the display corresponds to a particular carrier frequency of the tube. This reference baseline may be calibrated for the main oscillation frequency by tuning the signal oscillator A across the frequency range being swept. If the oscillator is swept over a range which is twice the intermediate frequency or greater, a double response will be obtained that allows for improved accuracy in the baseline calibration. The oscillation generator A may now be tuned over the band to be investigated to determine whether spurious oscillations exist. If a spurious oscillation is encountered, its frequency can be measured by adding or subtracting the intermediate frequency and the signal-generator frequency or by direct measurement with the wavemeter. The position of the signal on the oscilloscope screen indicates the voltage the oscillator under test requires to produce this spurious output. The corresponding main output frequency is known from the previous baseline calibration. However, care must be taken properly to differentiate a true spurious signal from spurious responses of the instruments. A mixer may easily generate harmonics of the CW signal and/or of the frequency of the oscillator under test and thus lead to responses at various frequencies. Use of the frequency meter and calculation of the harmonic relationships in observed responses will usually suffice to clarify this situation. When a spurious signal has been located, its amplitude can be measured on a comparative basis by introducing a simulated spurious signal via signal generator B.

For these measurements the main signal level is attenuated to the order of 1 mw, and the signal generator A is operated at a power output in the order of -30 dbm or less, to prevent overloading the receiver system. If the oscillator under study generates power greater than 1 watt, it will be convenient to sample its output by means of a directional coupler and provide a high-power matched load for the oscillator in the primary line. The sensitivity of the system at frequencies close to the carrier is limited by the selectivity of the IF amplifier.

14.2 Pulsed Oscillators

For pulsed oscillators, a spurious oscillation is any signal not part of the frequency spectrum normally associated with amplitude and frequency modulation of the carrier. Spurious oscillations may be generated on the rise or fall of the pulse, during a portion of the pulse, or during the complete pulse. They may repeat regularly on each pulse or may occur at random intervals.

Point-by-point measurements are presently the only commonly used method for detecting and measuring spurious outputs of pulsed oscillators.

14.2.1 Wavemeter Filter: A transmission-type cavity wavemeter may generally be employed in the circuit of Fig. 33 to investigate all forms of spurious oscillations. A sample of the RF output is coupled through the wavemeter to a detector and a pulse amplifier and is displayed on an oscilloscope. When the wavemeter is tuned to the main oscillation frequency, a pulse such as that shown in Fig. 33(a) is observed. If the tube fails to oscillate at this main frequency during any pulse, a baseline trace will be observed. As the wavemeter is tuned through its range, outputs at other frequencies may be observed. Spurious oscillations that are generated only on the rise and fall of the pulse will appear, as in Fig. 33(b). A spurious output pulse similar in shape to the main oscillation may appear at another

frequency and of much reduced amplitude. If information is desired as to the relative power produced at the main and spurious frequencies, the detector and indicator may be replaced by power-measuring equipment for direct measurement of this characteristic.

14.2.2 Beat-Frequency Technique: A sample of the RF output is applied to the circuit of Fig. 34. When the local-oscillator frequency is adjusted to bring the difference frequency between the local and the main oscillator frequencies within the amplifier pass band, a pulse-modulated signal such as that shown in Fig. 34(a) will appear on the screen. As the frequency of the local oscillator is brought closer to the frequency of the oscillator under study, the frequency of the beat signal will decrease until a zero beat condition will be observed. Spurious signals generated only on the rise and fall of the pulse will appear as in Fig. 34(b).

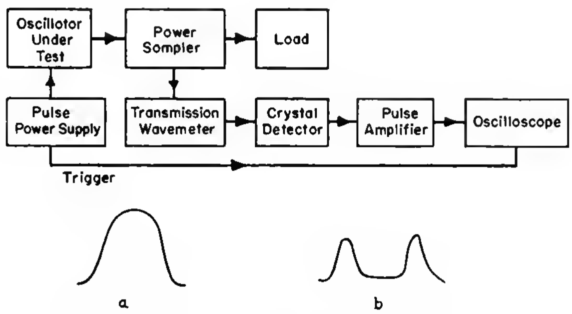


Fig. 33—Equipment for wavemeter-filter technique.

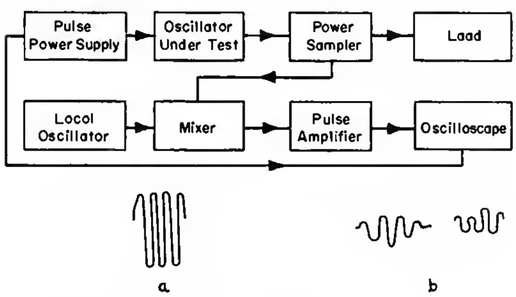


Fig. 34—Equipment for beat-frequency technique.

14.2.3 Spectrum Analyzer: A microwave spectrum analyzer[39] may conveniently be employed to observe spurious oscillations that exist continuously during each full pulse. The spectrum of the narrow pulses produced by spurious oscillations generated during the rise and fall of a pulse is generally too broad for meaningful observation. Moreover, because of the use of a swept frequency in a spectrum analyzer, spurious oscillations that occur randomly in time cannot usually be observed very reliably by this instrument.

15. Frequency Pushing

Frequency pushing is the change in operating frequency produced by a change in tube current. A numerical measure of frequency pushing is the pushing figure expressed in megacycles per ampere.

15.1 Static-Pushing Measurement

With the oscillator operating at a given input level, the frequency is measured by means of standard techniques. The input current is changed by a stated amount and the frequency is again measured. The frequency difference is the static pushing for the known change in input current. Static pushing includes some thermal effects, since there is a time interval between the two frequency measurements.

15.2 Dynamic-Pushing Measurement

Dynamic pushing, which avoids thermal effects, may be produced by varying the input current periodically. Usually the wave of current is a sine wave or a rectangular wave. Provision must be made to measure the peak difference in the changing current. The dynamic frequency change may be measured by a wavemeter and oscilloscope arrangement, by a spectrum analyzer, or by a receiver technique.

15.2.1 CW Oscillators: The arrangement shown in Fig. 35(a) may be used to measure the dynamic pushing of a CW oscillator by using sine-wave modulation. The variation of the input current is measured by either a meter or an oscilloscope. A sample of the RF energy is fed into a detector and a high-Q cavity wavemeter. The output of the detector is applied to the vertical in-

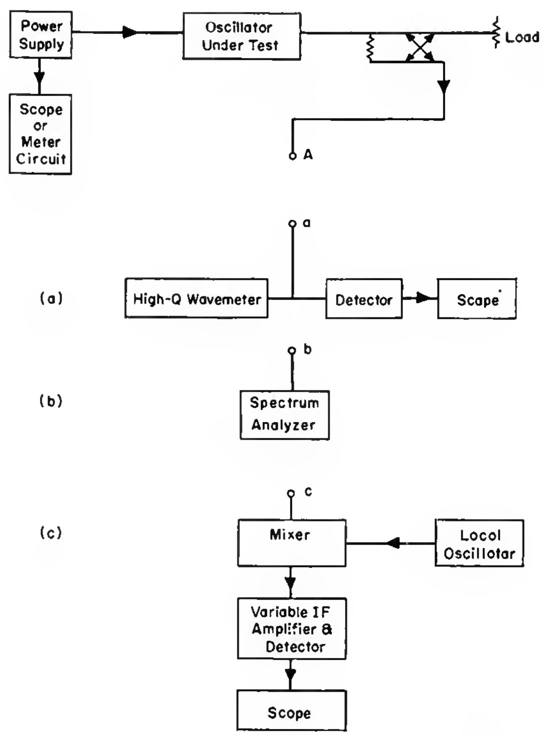


Fig. 35—Equipment for measuring pushing.

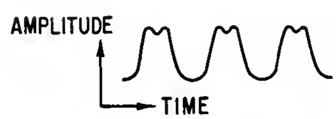


Fig. 36—Typical oscilloscope display for a CW oscillator.

put of an oscilloscope. The waveform shown in Fig. 36 indicates the display of the oscilloscope under typical conditions. The amplitude of the sine wave is a measure of the power output of the tube under modulated conditions. The pip results from the energy absorbed by the cavity wavemeter when it is tuned close to the tube frequency. As the wavemeter is tuned, the pip will progress from a crest to a valley and disappear. The difference in frequency between a crest and a valley measures the dynamic pushing of the tube for the change of input current employed. This presumes that the characteristic of frequency vs current is monotonic over the range of measurement. This will generally be true if the dynamic variation of the current is kept at a small value.

15.2.2 Pulsed Oscillators: The arrangement of Fig. 35(b) may be used to measure dynamic pushing of a pulsed oscillator by alternately pulsing it to different current levels. The change in the current may be measured by means of an oscilloscope, whose display will be similar to that shown in Fig. 37. A sample of the RF energy is applied to a spectrum analyzer. The resulting display is shown in Fig. 38. The difference in frequency between the two spectra measures the dynamic pushing of the tube for the change in current employed.

When dynamic pushing is small, this technique is not satisfactory since its accuracy depends on the separation of the two spectra. Under these conditions, the arrangement of Fig. 35(c) may be used. A sample of the RF energy is fed into a mixer where it is combined with the output of a CW oscillator. The IF signal from the mixer is fed into an IF amplifier whose intermediate frequency is variable and calibrated. (This may be a standard AM receiver that covers the 0.5- to 30-Mc band.) The detected output of the IF amplifier is applied to the vertical input of an oscilloscope. With the

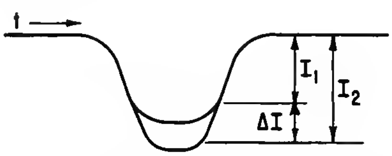


Fig. 37—Typical oscilloscope display showing change in current for a pulsed oscillator.

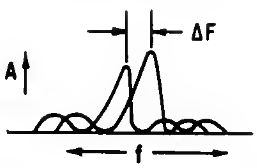


Fig. 38—Pulsed-oscillator display using spectrum analyzer.

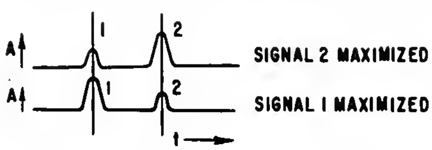


Fig. 39—Display obtained using circuit of Fig. 35(c).

oscilloscope synchronized to alternate pulses, a display similar to that shown in Fig. 39 will be seen. When the IF amplifier is tuned to the difference frequency between the large pulse and the local oscillator, one of the signals on the oscilloscope will be maximized. As the IF amplifier is tuned, this signal will decrease in amplitude and the other signal will increase until it is a maximum when the difference frequency between the small pulse and the local oscillator is reached. The dynamic pushing for the change in current employed is measured by tuning the IF amplifier to each peak and noting the frequency difference between the two settings.

C. Microwave Amplifiers

16. Matched Gain (Microwave Amplifier)

The matched gain of a microwave amplifier is defined as the ratio of 1) the RF power output into a reflectionless load to 2) the RF power incident at the input of the microwave amplifier. Two methods of measurement are used: the first depends on power measurements, and the second on attenuation measurements.

16.1 Direct Method

Referring to Fig. 40, a signal source supplies power through the power-measuring device to the amplifier under test. The output of the amplifier is connected through a second power-measuring device to a reflectionless termination. The power-measuring devices must be arranged to measure the power flowing in the forward direction. The gain of the amplifier under test is then the ratio of the measured output and input powers.

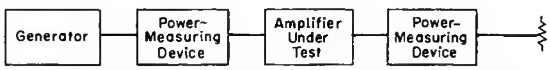


Fig. 40—Measurement of gain (direct method).

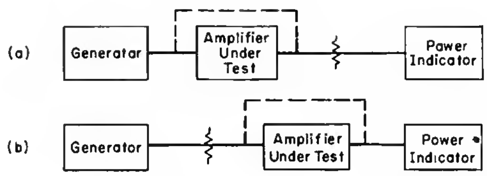


Fig. 41—Measurement of gain (indirect method).

16.2 Indirect Method

Referring to Fig. 41, a generator, calibrated variable attenuator, amplifier under test, and power indicator are connected and arranged in the order shown in either (a) or (b). The interconnecting transmission lines should be matched to the generator, attenuator, and power indicator. With the amplifier out of the circuit and replaced by a transmission line, a reference reading is established on the power indicator. With the amplifier in the circuit, the calibrated attenuator is adjusted to obtain again the reference reading on the power indicator. The gain is then equal to the change in attenuator read-

ings. Where limitations on the power output of the generator exist, arrangement (a) of Fig. 41 is preferable; where the limitation is the power-handling capability of the attenuator, (b) is preferable.

17. Input-Impedance Measurements

In microwave techniques it is conventional to use the voltage standing-wave ratio alone to indicate the degree of mismatch. However, by use of both the voltage standing-wave ratio S and the reference position of the standing-wave minimum expressed in electrical degrees θ, the impedance at the amplifier input terminals is defined. Assuming a lossless line between the detecting section and the microwave-amplifier input terminal, the input impedance Z is

$$Z = Z_0 \frac{1 - jS \tan \theta}{S - j \tan \theta},$$

where Z_0 is the characteristic impedance of the transmission line to the input of the microwave amplifier. Two methods will be given to determine S and one to determine θ. Since Z is a function of input level and other operating parameters of the amplifier, these tests should be performed under simulated operating conditions.

17.1 Measurement of Standing-Wave Ratio S

17.1.1 Standing-Wave Detector (Method A): Referring to Fig. 42, a signal source is connected to the amplifier through a standing-wave detector. The voltage standing-wave ratio S is measured.

17.1.2 Directional Coupler (Method B): Referring to Fig. 43, a signal source is connected to the amplifier through a directional coupler. The magnitude of the power (relative or absolute) in the forward and reflected waves is measured on the power indicators connected to the directional coupler. The reflection coefficient ρ, which is the square root of the ratio of the reflected to forward power, is related to the voltage standing-wave ratio S as follows:

$$S = \frac{1 + \rho}{1 - \rho}.$$

17.2 Measurement of the Reference Angle θ

The test arrangement shown in Fig. 42 is used in this measurement. A reference position of a standing-wave minimum is established by short-circuiting the amplifier input terminals. With the short circuit removed, the distance between the reference position and the nearest minimum in the direction of the generator is measured. This distance expressed in electrical degrees is the reference angle θ

18. Output-Impedance Measurements

For the same reasons as given for input-impedance measurements, the output impedance of a microwave amplifier is determined by use of the standing-wave ratio S and the reference angle θ. Assuming lossless transmission lines, the output impedance Z is

$$Z = Z_0 \frac{1 - jS \tan \theta}{S - j \tan \theta},$$

where Z_0 is the characteristic impedance of the transmission line connected to the output of the amplifier. Methods will be given to determine S and θ: These tests should be performed under simulated operating conditions.

18.1 Measurement of Standing-Wave Ratio S

Referring to Fig. 44, a signal source is connected to the amplifier under test. This signal generator should be matched to the interconnecting transmission line. At the output of the amplifier, a directional coupler, standing-wave introducer, and matched termination are connected in series. A power indicator is connected to the directional coupler to indicate forward-going power. The standing-wave introducer is adjusted to produce a known reflection coefficient ρ_T. A maximum reading $R_{\max}$ and a minimum reading $R_{\min}$ are obtained on the power indicator by movement of the standing-wave introducer. The square root of the ratio $R_{\max}$ to $R_{\min}$ is the system standing-wave ratio $\overline{S}$. The relation between $\overline{S}$, the reflection coefficient of the amplifier ρ_A and the termination reflection coefficient ρ_T is as follows:

$$\overline{S} = \frac{1 + \rho_A \rho_T}{1 - \rho_A \rho_T}.$$

Since ρ_T and $\overline{S}$ are known, ρ_A may be determined. The amplifier-output SWR is then

$$\overline{S} = \frac{1 + \rho_A}{1 - \rho_A}.$$

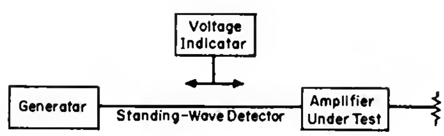


Fig. 42—Measurement of SWR and reference angle.

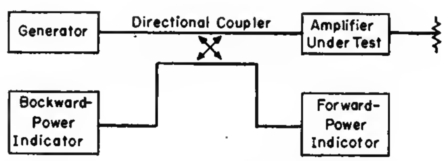


Fig. 43—Measurement of SWR using directional coupler.

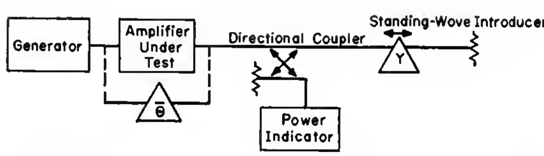


Fig. 44—Measurement of SWR and reference angle.

18.2 Measurement of Reference Angle θ

With the amplifier in the circuit as shown in Fig. 44, the movable standing-wave introducer is adjusted to produce a minimum of the power indicator, and its position recorded. A fixed standing-wave introducer with known reflection-coefficient angle $\bar{\theta}$ is substituted for the amplifier under test. The movable standing-wave introducer is shifted toward the generator until the next minimum is obtained. The distance between the two minima expressed in electrical degrees is called ψ. The reference angle θ is then

$$\theta = 2\psi + \bar{\theta}.$$

19. Measurement of Amplifier Bandwidth

The amplifier bandwidth is the frequency interval over which a specified characteristic of the amplifier varies within prescribed limits. The characteristic may be gain, noise figure, power output, etc. Conventionally the amplifier bandwidth refers to gain bandwidth and can be determined from the measurement of gain, as described in Section 16. A sufficient number of measurements should be taken to establish clearly the parameter-vs-frequency curve. These measurements are done ideally with a swept-frequency source of constant power output.

20. Measurement of Amplifier Loss

20.1 Circuit Insertion Loss

The circuit insertion loss of a microwave amplifier is defined as the ratio of the RF power input to the RF power output with the amplifier under nonoperating conditions. The procedure for this measurement is identical with that used in the gain measurements described in Section 16.

20.2 Backward Loss

The backward loss of a microwave amplifier is defined as the ratio of the RF power accepted by the output terminals of the amplifier to that appearing at the input terminals. When the amplifier is under nonoperating conditions, the backward loss is identical with the circuit insertion loss; however, under operating conditions the backward loss will not in general be the same as the circuit insertion loss, and will be a function of the applied potentials.

The procedure for this measurement is again identical with that used in the gain measurement described in Section 16, except that the signal generator is connected to the amplifier output terminals.

21. Phase Measurements

This section will include phase measurements and a time-delay measurement for a short pulse under two sets of conditions. In Section 21.1 the frequency is held constant and the variations in the phase of the output signal are measured as a function of the various amplifier operating conditions. In Section 21.2 the amplifier conditions are held constant, and the phase change of the output signal as a function of the frequency is measured. In Section 21.3 the amplifier conditions are held constant at the values used in Section 21.2, a short pulse centered at the frequency used in Section 21.1 is passed through the amplifier, and the delay time through the amplifier is measured.

21.1 Fixed-Frequency Tests

Referring to Fig. 45(a), a signal generator, a directional coupler, and an amplifier are connected in cascade. The output of the amplifier is connected to a directional coupler and load in cascade. The forward-going power sampled by the directional couplers is applied to the two ends of a standing-wave detector. For maximum measurement accuracy, the sampled powers arriving at the voltage indicator should be of nearly equal magnitude.

The measurement procedure is as follows: 1) a reference position of the traveling probe is established by adjusting the position of the probe for a minimum reading on the voltage indicator; 2) the shift of the probe position in electrical degrees from the reference position to a position where the voltage indicator again shows a minimum is measured as a function of the change in operating conditions under consideration, for example,

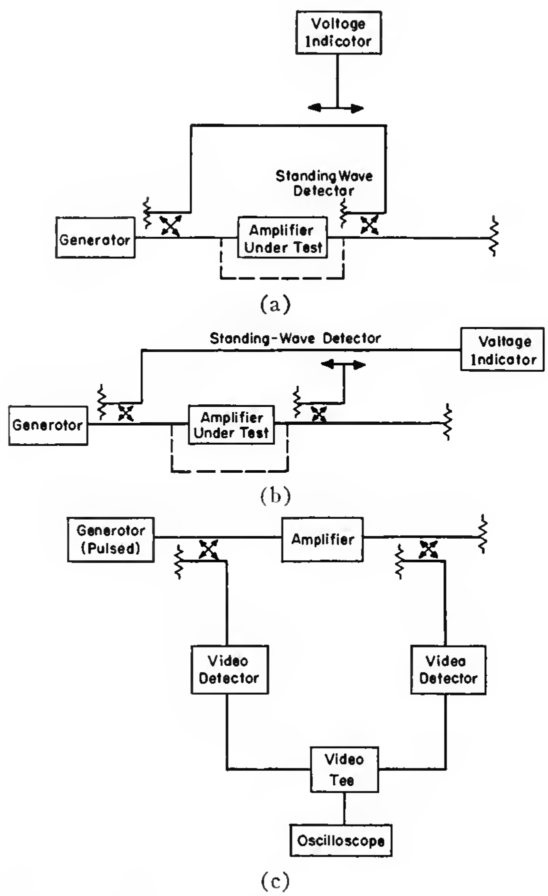


Fig. 45—System for phase measurements. (c) Measurement of time delay.

electrode potentials, electron current, or input RF drive.

The phase shift of the output signal in electrical degrees is twice the shift of the probe position.

An alternate method of measuring phase shift of the output signal is shown in Fig. 45(b). In this technique, the output signal sampled by the directional coupler goes to the probe of the standing-wave detector, and the voltage indicator is connected to one end of the standing-wave detector. In all other respects, the measuring setup is identical with Fig. 45(a). The measurement procedure is as before, except that the phase shift of the output signal in electrical degrees will be equal to the shift of the probe position.

21.2 Variable-Frequency Test

Using the arrangement in Section 21.1, but with the amplifier out of the circuit and the input and output directional couplers joined, the change in phase shift vs frequency of the circuit is determined, as described in Section 21.1.

With the amplifier replaced in the circuit, a second measurement of change in phase shift vs frequency is made.

The change in phase shift of the output signal of the amplifier as a function of frequency is then the difference between the two measurements.

21.3 Time-Delay Measurement

Referring to Fig. 45(c), a generator producing RF pulses is connected to the amplifier. The input and output are sampled by directional couplers. The input and output RF pulses are detected and introduced into a matched video tee junction. The combined pulses are amplified and displayed on a wide-band oscilloscope with a calibrated sweep to determine time delay.

The time delay determined by the pulse spacing must be corrected for any difference in delay external to the amplifier under test.

22. Noise Factor, Noise Figure

This measurement is usually of interest with regard to low-level amplifiers, particularly when signals approaching noise level are being considered. Noise figure may also be used to calculate the SNR of high-level linear systems.

Noise figure may be defined as follows: The noise figure of a linear system "at a selected input frequency is the ratio of (1) the total noise power per unit bandwidth (at a corresponding output frequency) available at the output terminals, to (2) the portion thereof engendered at the input frequency by the input termination, whose noise temperature is standard (290°K) at all frequencies."[44]

The method of measuring the noise figure of a microwave amplifier when using a noise source will be discussed in the following paragraphs.

As shown in Fig. 46, a noise source is connected to the amplifier through a band-pass filter.[45] The output of the amplifier is connected through a second band-pass filter to a relative-power-measuring device. The relative-power-measuring device is in general a superheterodyne receiver in which a calibrated variable attenuator has been inserted in the band-pass-filter section. Relative power is read from this calibrated attenuator; the signal level at the receiver detector is held constant.

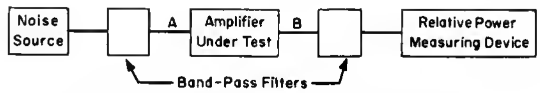


Fig. 46—Measurement of noise figure.

The band-pass filters are tuned to the identical center frequency and have filter characteristics such that spurious or image responses are eliminated at the power-measuring device. The filter at the input to the amplifier may also serve to prevent a wide-band noise source from overloading a wide-band low-level amplifier. The power-measuring device should have a bandwidth narrower than the filters and the amplifier under test.

To establish the noise figure of the power-measuring device, the amplifier is taken out of the circuit and A is connected to B directly. With the noise source inoperative, a reading P_1 is registered by the power indicator. With the noise source on, a second reading P_2 is taken. If the power from the noise source is b times $kT\Delta f$, the power-measuring device noise figure F_R at point B is then

$$F_R = \frac{P_1(b-1)}{(P_2 - P_1)a},$$

where a is the loss between the noise source and point A expressed as a numerical ratio ($a \geq 1$), k is Boltzmann's constant, T is the ambient temperature, and Δf is the bandwidth of the power-measuring device.

With the amplifier in the circuit, the same procedure is followed to establish the over-all noise figure F_T. If the gain G of the amplifier is known, the amplifier noise figure F_A is given by

$$F_A = F_T - \frac{F_R - 1}{G}.$$

23. Carrier-to-Noise Fluctuations

23.1 Amplitude Fluctuations

This measurement is usually of interest in the case of high-level amplifiers. The carrier-to-noise test is a measure of the ratio of carrier-amplitude fluctuations to average carrier amplitude. These fluctuations arise

[44] "Standards on electron devices: methods of measuring noise," Proc. IRE, vol. 41, p. 896; July, 1953.

[45] The noise source and filter should be matched to the interconnecting transmission line. The noise source should be matched both when switched on or off.

from such sources as shot noise, thermal noise, electrode-voltage fluctuations, and spurious oscillations within the tube. The noise-power spectrum is not ordinarily constant with frequency. For this reason, the results of this test cannot be reduced to the noise power expressed on a per cycle basis.

In the test arrangement shown in Fig. 47 a signal source and the amplifier under test are connected to a matched termination through a directional coupler. The output from the coupler is fed through a calibrated attenuator to a suitable detecting circuit. The power level at the detector must be such that the detection is linear over the range of the fluctuations to be measured. The output of the detector is connected to an indicating device of appropriate video bandwidth.

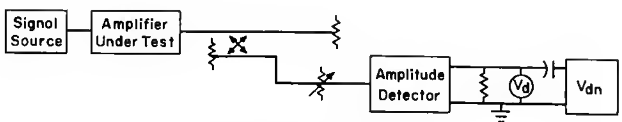


Fig. 47—Measurement of amplitude fluctuations.

Calibration of Detector: An initial calibration is necessary in order to determine the voltage change produced across the resistor in the detector circuit by a given small change in carrier amplitude. The deviation ΔV_d in direct voltage across the resistor R is measured when the calibrated attenuator is adjusted to change the carrier level by a known amount ΔV_c. This change in level must be sufficiently small to assure linear operation of the detector.

Note: In this procedure the ratio $\Delta V_c/V_c$ is obtained from the calibrated attenuator.

By means of a band-limited voltage-measuring device, the rms value of the alternating component V_{dn} of the voltage across the resistor R in the detector circuit is measured. The bandwidth and center frequency of the voltage-measuring device are chosen to coincide with the intended operating conditions.

The SNR expressed in decibels, is

$$S/N = 20 \log_{10} \frac{V_d}{V_{dn}} \left[\frac{(\Delta V_d/V_d)}{(\Delta V_c/V_c)} \right].$$

This measurement assumes that the source and the detecting system are essentially noise free.

Caution: dc voltmeter may affect the video circuit impedance.

23.2 Phase Fluctuations

This measurement is usually of interest in the case of high-level amplifiers. The test will provide a measure of carrier-phase fluctuations. These fluctuations arise from such sources as shot noise, thermal noise, electrode-voltage fluctuations, or spurious oscillations within the tube. The noise spectrum is not ordinarily constant with frequency or level of operation. For this reason, this test cannot be reduced to a completely general measurement of the noise expressed on a per cycle basis.

The test arrangement shown in Fig. 48 consists of a conventional phase bridge with two hybrid junctions. One arm contains the amplifier under test with appropriate attenuators and power-monitoring devices, and the other arm contains an adjustable phase standard and an attenuator. The side arms of the second hybrid are connected to crystal detectors of opposite polarity which feed into equal arms of a dc bridge. The bridge output is fed to a band-limited voltage amplifier and indicator, as well as to a dc voltmeter. One side of the dc bridge is grounded for crystal-current return.

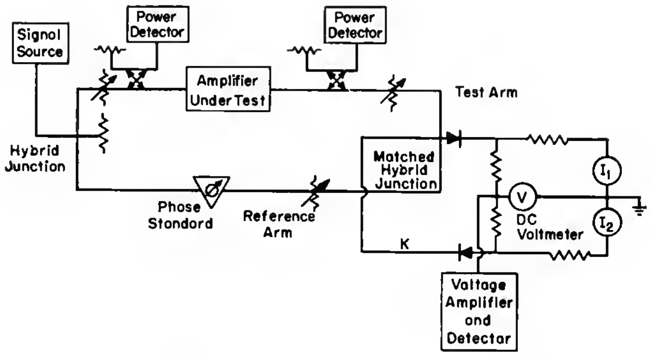


Fig. 48—Measurement of phase fluctuations.

Adjustments:

1) For the first adjustment, the signal applied to the reference arm must be very small. In order to check balance in the second hybrid and the detecting system, the output from the amplifier under test (operating at the desired measurement level) is suitably attenuated and fed into the test arm of the second hybrid. Under these conditions, the readings of the two current meters should be equal. Inequality of readings indicates an unbalance in the detecting system which must be corrected.

2) With the signal in the test arm zero, a signal is fed into the reference arm and adjusted so that the current meters indicate currents equal to those previously noted in Step 1). These two steps insure equal levels of the signal inputs to the test arm and reference arm of the hybrid.

3) With signals fed into both the test arm and the reference arm, and set at levels 3 db below those determined in Steps 1) and 2) (to maintain the total level at the phase-detector constant), the phase shifter in the reference arm is adjusted so that the dc voltmeter across the bridge output reads zero. This insures that the signals at the inputs to the second hybrid are in quadrature.

Calibration: The dc voltmeter reading is proportional to phase deviation in either arm for small phase deviations in the balance condition. A direct calibration is obtained by recording voltmeter readings for small adjustments of the phase standard in the reference line.

This procedure gives the system sensitivity in volts per degree of phase change.

Measurement: Random phase fluctuations in the amplifier under test will appear as voltage fluctuations at the output of the bridge circuit. Under the balance conditions noted above, the rms value of these fluctuations is measured by use of the band-limited voltage-measuring device. The reading in rms volts may be converted to rms phase flucutations by means of the calibration made in this section. Peak-to-peak measurements may be made in a similar manner.[46]

Severe errors in measurement may occur if care is not taken to observe the following:

1) All microwave components must be well matched.
2) The second hybrid must be well balanced.
3) The signal source must be frequency-stabilized.
4) Matched crystal pairs must be used.
5) The total phase delay of the reference arm between the first and second hybrid arm must be approximately equal electrically to that of the test arm.
6) The phase-standard calibration is the primary calibration involved.
7) For certain tubes, large phase deviations may result from power-supply ripple, heater-current fluctuations, etc. Care must be taken to separate these effects from true noise fluctuations.

24. Frequency Range

Amplifier frequency range is the frequency interval over which any particular characteristic of the amplifier can be held within prescribed limits by adjusting the operating parameters. The characteristics involved may be gain, noise figure, power output, etc. The parameters adjusted may be anode-voltage, field, or other operating parameters. This determination of the frequency range may be made with respect to gain, noise figure, power output, etc. Conventionally, the amplifier frequency range refers to gain frequency range. A sufficient number of measurements should be made to establish clearly the curve of the parameter vs frequency.

25. Amplifier Power Output

The power output of a microwave amplifier is dependent on the direct operating voltages applied to the various electrodes, the external load impedances, the level of applied input power, and on additional external parameters, such as magnetic-field strength. This power output is composed of the sum of the fundamental and all harmonic powers.

25.1 Amplifier Fundamental Power Output

The amplifier fundamental power output is defined as the output power at the fundamental frequency of the applied signal. Throughout the test it will be assumed that the source is supplying a CW signal of negligible harmonic content.

As shown in Fig. 49, a signal source is connected through a calibrated attenuator and a power-measuring device to the amplifier under test. The output of the amplifier is fed through a calibrated attenuator to a filter. The filter is chosen such that the harmonic transmission is sufficiently small relative to the fundamental transmission.

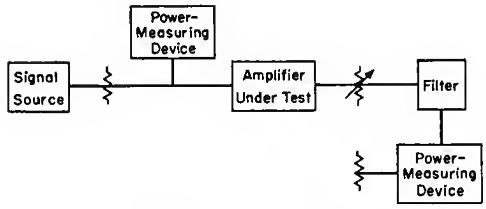


Fig. 49—Measurement of fundamental or harmonic power output.

The output of the filter is connected through a power-measuring device to a suitable termination. The power-measuring devices must be arranged to measure the power in the forward direction only.

By means of this test arrangement, the fundamental output power for any desired set of operating conditions may be measured.

25.2 Harmonic Power Output

For this measurement, it is necessary to alter the filter shown in Fig. 49 so that only power at the particular harmonic frequency will pass. With this altered arrangement, the power output at any given harmonic can be measured for any desired set of operating conditions. In general, harmonic power is expressed as a percentage of fundamental power. This percentage can be computed directly from the values of fundamental frequency.

Precautions:

1) The signal source must be sufficiently well matched so that no reflections of harmonic power will take place in the input line.

2) Since harmonics may be propagated in more than one mode on a given structure, care should be taken that the mode transmission characteristics of all components and detectors used are known in order to insure that the total harmonic power is measured.

26. Conditional Oscillations

Under certain load or source impedance conditions, it is possible that oscillations may occur in a microwave amplifier. Such oscillations are herein called *conditional oscillations.*

As shown in Fig. 50, a standing-wave introducer and termination are connected to the input and output of the amplifier under test. In addition, suitable power-measuring devices are connected in the input and output lines.

The test is accomplished by observing the power

[46] D. A. Alsberg and D. Leed, "Phase and transmission measurement," *Bell Sys. Tech. J.*, vol. 28, pp. 221–238; April, 1949.

detector as the input and output impedances are varied over any desired range. The operating parameters of the amplifier should be varied (swept) over prescribed ranges during the test. Oscillations will show up as indications of power output.

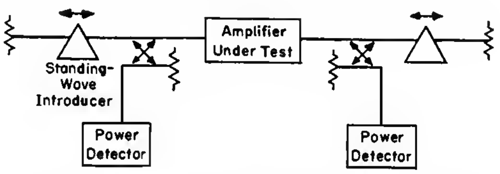


Fig. 50—Measurement of conditional oscillations from an amplifier.

Precautions:

1) The bandwidth of the directional coupler and the power detector must be sufficient to permit detection of oscillations in the frequency range of the amplifier.

2) Care must be taken to differentiate between noise output and oscillation output.

27. Intermodulation

Intermodulation is defined as "the modulation of the components of a complex wave by each other in a nonlinear system."[47] Six distinct tests for the major types of intermodulation will be described. The first four tests are concerned with testing for the production of new side frequencies or alteration of the relative phase and amplitude of existing side frequencies because of nonlinear effects when a single modulated carrier is passed through a microwave amplifier. The last two tests concern the interaction between two or more independent carriers, which may or may not be modulated.

The six tests are:

1) Amplitude distortion.
2) Phase-to-amplitude conversion.
3) Phase distortion.
4) Amplitude-to-phase conversion.
5) Multiple-signal intermodulation (frequency-conversion effect).
6) Cross modulation.

The foregoing tests are concerned primarily with nonlinear effects. A signal modulated in any fashion can suffer distortion by virtue of the change in the relative amplitude and phase of the side frequencies in a linear amplifier or network without the creation of new side frequencies. These distortions can be calculated from the static tests of gain and phase shift vs frequency (refer to Sections 16 and 21.2). This type of distortion can be equalized by passive networks. Dynamic measurements of this type of distortion can be made by methods to be described. In addition, nonlinear effects can also change relative phase and amplitude of side frequencies and, in general, phase and amplitude distortion from both sources are present simultaneously in a nonlinear amplifier.

[47] "Standards on receivers: definition of terms, 1952," 52 IRE 17.S1, Proc. IRE, vol. 40, p. 1683; December, 1952.

In the following tests a signal source that is sinusoidally modulated in either amplitude or phase alone is often called for. Such a source, if available, makes the interpretation of the measurements comparatively simple. If such a source is unavailable, the additional distortions due to the signal source must be taken into account. Alternately, it may be simpler to use a square-wave modulation of the signal source. The implementation of the test is then simpler, but the interpretation in some cases is more complex. Where a particular application is intended for the amplifier under test, it is advisable to use a signal and a receiver similar to those to be used in the intended application. In any of these instances, however, the basic principles are the same as those outlined in the following tests.

27.1 Amplitude Distortion

A measurement of amplitude distortion can be made by means of the test arrangement of Fig. 51. The amplitude-modulated output of an RF signal source is fed through a variable attenuator to the input of the amplifier under test. A portion of the amplifier output is coupled through a directional coupler, a variable attenuator, and an RF filter (capable of removing RF harmonics) to a suitable linear detector. The output terminal of the detector is connected to a harmonic analyzer capable of measuring individual harmonics of the modulating signal.

With the signal generator modulated at the desired modulation percentage and the amplifier under test operating under the desired conditions, the various harmonics of the modulating signal appearing in the detector output are measured and are expressed as a percentage of the modulating-frequency component.

The detector should be operated at the same level for measurement of the fundamental and for each harmonic component. Every precaution should be taken to assure that the detector is not introducing additional distortion. It is important that the amplitude linearity be checked initially by bypassing the amplifier under test. The amplitude of each harmonic present in the driving source must be small as compared with the corresponding harmonic present in the amplifier output. It should be noted that amplitude distortion is, in general, a function of the input-carrier level, the modulation percentage of the carrier, and the operating conditions (electrode voltage, etc.), of the amplifier.

Fundamentally, the technique of making this measurement is identical with that used at audio or video frequencies.

It may be desirable to modulate the carrier with a complex waveform and search for intermodulation components.

27.2 Phase-to-Amplitude Conversion

This test may be made with the same equipment as that shown in Fig. 51 but the signal source is phase-or-frequency-modulated, instead of amplitude-modulated.

Any amplitude modulation caused by the amplifier under test can be measured at the output of the detector, the harmonic analyzer being used to measure the relative amplitudes of the various harmonic components. The result can be expressed for the fundamental and each harmonic as a percentage modulation of the carrier for a specified phase or frequency deviation. It is important that the inherent amplitude modulation of the signal source be negligible.

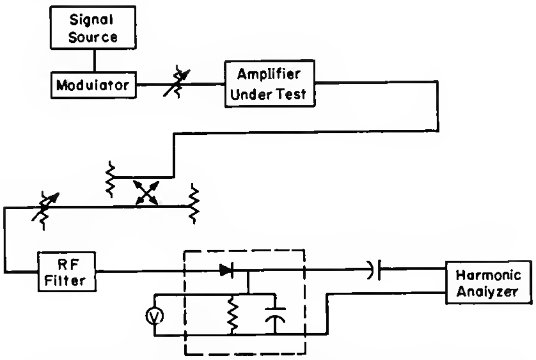


Fig. 51—Arrangement of equipment for amplitude-distortion measurement.

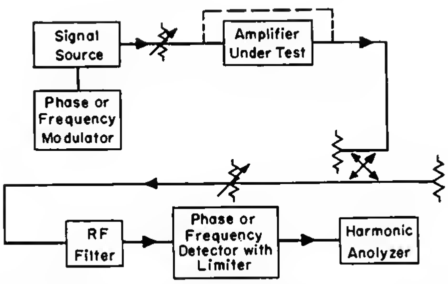


Fig. 52—Arrangement of equipment for measurement of phase distortion.

27.3 Phase Distortion

A measurement of phase distortion can be made by means of the test arrangement of Fig. 52. The phase-or-frequency-modulated output of an RF signal source is fed through a variable attenuator to the amplifier under test. The output of the amplifier is fed through a directional coupler, a variable attenuator and RF filter (capable of removing harmonics of the carrier) to a suitable phase or frequency detector, which should include a limiter. The output of the detector is fed to a harmonic analyzer capable of measuring individual harmonics. With the signal generator modulated in the desired manner, the various harmonics of the modulating signal appearing in the detector output are measured and expressed as a percentage of the modulating-frequency component.

The detector should be operated at the same level as above, and the limiter must remove amplitude variation without introducing additional phase distortion. Amplitude variations can result from phase-to-amplitude conversion in the amplifier under test. (Refer to Section 27.2, Phase-to-Amplitude Conversion.)

Incidental amplitude modulation of the signal source must be small, since it can cause amplitude-to-phase conversion in the amplifier under test and thus give rise to additional phase-modulation components of sufficient amplitude to affect the accuracy of the test. (Refer to Section 27.4, Amplitude-to-Phase Conversion.)

In phase-distortion measurements special care must be taken to insure that the amplifier is inserted into a line that is essentially reflectionless in both directions to minimize long-line effects. It is important that the system phase linearity be checked initially by bypassing the amplifier under test. Under these circumstances, the amplitude of each distortion component at the output of the phase detector should be small compared with the corresponding distortion component produced by the amplifier under test.

27.4 Amplitude-to-Phase Conversion

In this test an amplitude-modulated signal is transmitted through the amplifier under test. The phase modulation caused by the amplifier is measured by a suitable detector. This test may be made with the same equipment as is shown in Fig. 52, except that the signal source is now amplitude-modulated. The equipment is used essentially as described in Section 27.3, except that the phase modulation measured is now that caused by amplitude-to-phase conversion in the amplifier. The result is usually expressed as the rms phase deviation for a given AM percentage. At the signal levels at which appreciable amplitude-to-phase conversion occurs, the amplifier may also deliver appreciable power at harmonics of the carrier frequency. It is important that the harmonic components be eliminated from the bridge circuit. This can be done by using an RF filter at the output of the amplifier under test. The filter must have sufficient bandwidth to pass all of the sidebands created by the phase-modulation process. The phase characteristic of the filter should be essentially linear over the filter bandwidth.[48]

It should be noted that, even though the input signal is sinusoidally modulated, harmonics of the modulating frequency may appear in the output as a result of distortion in the amplitude-to-phase conversion process. If these harmonics are of interest, they can be measured by using a harmonic analyzer as the output indicator.

27.5 Multisignal Intermodulation (Frequency-Conversion Effect)

When two unmodulated carriers having frequencies f_1 and f_2 are passed through a microwave amplifier, nonlinear effects produce new frequencies equal to $mf_1 \pm nf_2$, where m and n are integers. When these spurious frequencies fall within the band of interest of the intended use, they may be of concern. The amplitudes of these spurious frequencies are dependent upon the amplitudes

[48] C. F. Augustine and A. Slocum, "Six kmc phase measurement system for traveling wave tubes," IRE TRANS. ON INSTRUMENTATION, vol. PGI-4, pp. 145–154; October, 1955.

and frequencies of the input signals.

Furthermore, as a result of the same nonlinear effects, the gain at a particular frequency is affected by the amplitude and frequency of a second independent signal. When more than two signals are passed through the amplifier, the same phenomenon takes place. A two-signal test will be described. It can be readily extended to as many signals as desired.

For this test, two signals of frequencies f_1 and f_2, and of known amplitudes, are simultaneously fed to the amplifier under test. A method of doing this is shown in Fig. 53. The output of the amplifier is sampled by a tunable narrow-band receiver and indicator. By means of the receiver, the amplitudes of all of the output signals in the frequency band of interest are measured. The results of such a measurement may be conveniently displayed in the form of a graph, an example of which is shown as Fig. 54. Here, the input power at f_2 has been held constant, and the input power at f_1 has been varied over a wide range. The components $2f_1-f_2$ and $2f_2-f_1$ are spurious signals generated in the amplifier. A typical measure of the amplifier performance is the ratio, in decibels, of the power output at a specified spurious signal frequency to the power output at one of the input frequencies.

In order that the amplitude and frequency of the two signal sources may be adjusted independently, care must be taken to insure that the two sources do not interact.

The receiver must have adequate selectivity and dynamic range to measure correctly each separate signal.

The sensitivity of the receiver and the attenuation of the coupling system must be known over the band of interest.

The receiver must be operated at such a level that any intermodulation generated in the mixer is negligible.

The system can be checked by bypassing the amplifier under test.

27.6 Cross Modulation

In this test two carriers of frequencies f_1 and f_2 are transmitted simultaneously through the amplifier under test. The carrier f_1 is modulated in amplitude or phase, and the amount of amplitude or phase modulation appearing on the second carrier f_2 is measured. When carrier f_1 is modulated only in either amplitude or phase, both amplitude and phase modulation may appear on carrier f_2. This test can readily be extended to more than two signals.

As shown by Fig. 55, a modulated signal source f_1 and an unmodulated signal source f_2 are connected through attenuators and directional couplers to the amplifier under test. The output of the amplifier is connected to a matched termination through a directional coupler. The output of the coupler is fed through a variable attenuator to a microwave receiver. The receiver must incorporate an RF-level indicator as well as a suitable detector for the type of modulation of interest.

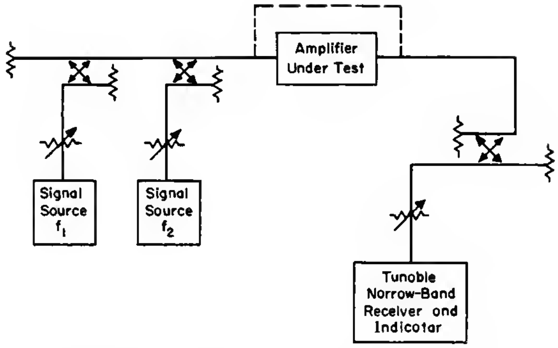


Fig. 53—Arrangement for cross-modulation test.

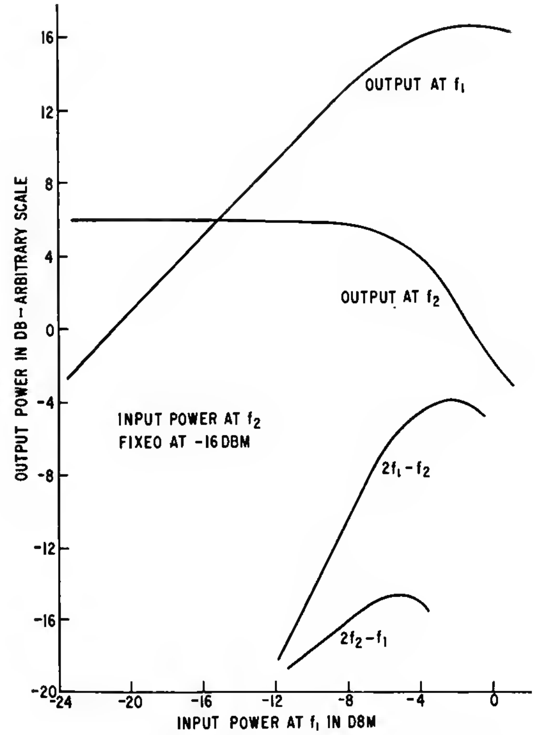


Fig. 54—Multisignal-modulation typical result of two-signal measurement.

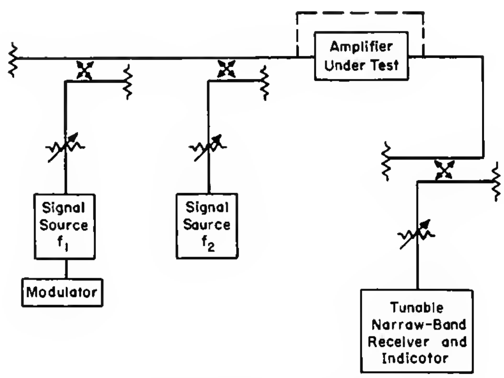


Fig. 55—Arrangement for measurement of multisignal intermodulation.

With the receiver tuned to detect carrier f_1, the average RF level and the modulation level are observed. The receiver is then tuned to carrier f_2 and the average RF level adjusted by the attenuator in the receiver line to be equal to that previously observed. The modulation appearing on the second carrier is then measured. The result obtained may be a function of the power inputs to the amplifier under test and of the type and amount of modulation.

Precautions: All transmission lines are suitably matched to prevent interaction of the signal sources. The inherent cross modulation of the system should be checked by bypassing the amplifier under test and repeating the measurements. A second detector which is essentially linear over the range of modulation must be used.

28. Modulation Characteristics

By variation of suitable electrode potentials, the amplitude and phase of the output signal of a microwave amplifier can be changed. This property can be used to modulate a signal. Usually by choice of a proper modulation electrode and correct operating parameters, either amplitude or phase changes can be made predominant. Generally, however, residual modulation of the other type is also present. Modulation characteristics usually of interest are:

Amplitude-modulation curve: (The output amplitude as a function of an operating parameter.)

Phase-modulation curve: (The output phase as a function of an operating parameter.)

Amplitude-modulation sensitivity: (The slope of the amplitude-modulation curve at a stated point.)

Phase-modulation sensitivity: (The slope of the phase-modulation curve at a stated point.)

Nonlinear modulation characteristics: (Distortion in the modulation process that occurs when the amplitude of the modulating signal becomes sufficiently large. This distortion causes modulation at harmonics of the modulating signal as well as at its fundamental frequency. The degree of modulation occurring at the modulating frequency and at its harmonics is determined as a function of the amplitude of a sinusoidal modulating signal.)

Residual modulation: (The amount of phase modulation present when amplitude modulation is desired, or conversely the amount of amplitude modulation present when phase modulation is desired.)

Modulation impedance: (The impedance presented to the modulator by the tube under consideration. It should be noted that in some cases this impedance may be nonlinear, and that harmonic components of current may therefore be drawn from the modulator. The modulation impedance is in general a function of the modulating frequency.)

In principle the amplitude- or phase-modulation characteristics may be measured by changing the operating parameter of interest while maintaining the incident signal input power constant and measuring the change in output power, as in Section 16, or the change in output phase, as in Section 21. However, in some cases, because of thermal effects in the amplifier, the amplitude- or phase-modulation curves measured dynamically may be different from those measured point-by-point. Furthermore, a dynamic measurement lends itself more readily to the practical determination of the modulation sensitivities, nonlinear modulation characteristics, and residual modulation than does a point-by-point measurement. For these reasons modulation characteristics are frequently measured by dynamic methods.

For dynamic measurements of modulation characteristics, adaptations of tests described in Section 27 may be used.

To measure the amount of amplitude modulation, equipment similar to that shown in Fig. 51 and described in Section 27.1 is employed, but the amplifier under test, rather than the signal source, is modulated. To measure the AM sensitivity, a small modulating signal is applied to the amplifier under test and the resulting modulation is measured. To measure the nonlinear modulation characteristics, a modulating signal of the desired amplitude is applied to the amplifier under test, and the fundamental and harmonics of the modulating frequency are measured by the harmonic analyzer. In general, the ratio of the amplitudes of the harmonics to that of the fundamental is a function of the amplitude of the modulating signal. The equipment of Fig. 51 may also be used to measure the residual amplitude modulation when the modulating signal produces mainly phase modulation.

To measure the amount of phase modulation, equipment similar to that shown in Fig. 52 and described in Section 27.3 is employed, but the amplifier under test, rather than the signal source, is modulated. The phase-modulation sensitivity and the nonlinear phase-modulation characteristics are then determined in a manner analogous to the procedure outlined in the preceding paragraph for corresponding AM characteristics. To measure the residual phase modulation in the presence of amplitude modulation, the equipment shown in Fig. 50 and described in Sections 23.2 and 27.4 is employed, but the amplifier under test, rather than the signal source, is modulated. The use of the phase bridge of Fig. 48 rather than the equipment in Fig. 52 is preferred in this case because large amplitude modulation may cause sufficient conversion of amplitude modulation to phase modulation in the limiters in Fig. 52 to affect the phase modulation measured at the output.

The modulation impedance is measured by conventional LF techniques. In most cases the methods are the same as those used in measuring the input impedances of active devices.

29. Testing of Microwave Amplifiers under Pulse Conditions

For the purpose of the following discussion, it is desirable to distinguish between three categories of pulse

amplification. RF amplifiers providing pulsed outputs may be divided into the following three categories:

1) Amplifiers in which electrode voltages are pulsed, but the input signal is CW.
2) Amplifiers in which the input signal is pulsed, but electrode voltages are not pulsed.
3) Amplifiers in which electrode voltages and input signals are both pulsed.

Most of the tests previously discussed may be modified and applied to all three categories. Such modifications commonly involve either the measurement of average power and the computation of peak power from a knowledge of duty factor or the measurement of peak power by comparison techniques.[49]

In addition to the tests previously described, there are several tests for specific amplifier characteristics. These are described below.

29.1 Pulse Shape and Spectrum

Microwave amplifiers can cause a change in the shape and spectrum of RF pulses both because of their bandwidth limitations and because of nonlinear effects. If the RF input is pulsed (categories 2 and 3), the output pulse may differ from the input pulse in shape and spectral distribution. Such differences can be determined by simultaneously displaying the RF input and the output pulses or spectra, as shown an Fig. 56.

If the electrode voltages are also pulsed (cateory 3), care must be taken to separate effects caused by microwave-amplifier distortion from those caused by modulator-waveform distortion. (The effects caused by modulator-waveform distortion are particularly important when the input RF pulse starts during or before the electrode-voltage pulse.)

With a CW input signal (category 1), where it is required to measure the shape and spectrum of the output pulse and to compare the pulse shape with the shape of the electrode-voltage pulse delivered by the modulator, the arrangement shown in Fig. 57 is used. It is desirable to observe the time delay of output RF pulses with respect to electrode-voltage pulses.

29.2 Pulse-to-Pulse Phase Coherence

29.2.1 Phase Jitter: Phase jitter is the variation in phase delay introduced by the amplifier during a pulse or between successive pulses. Phase jitter can be measured by the equipment shown in Fig. 58. Essentially, an RF bridge circuit is used to compare the RF phases of the input and output signals during the pulse interval. Any variation in the output phase from pulse to pulse, or during the period of a pulse, will produce a video output of the phase bridge. This output is a measure of the phase jitter and can be observed by means of a voltmeter or an oscilloscope. The phase jitter caused by the modulator must be small enough to insure that it does not influence the measurement.

[49] Section 5.2, Pulse-Power Measurements.

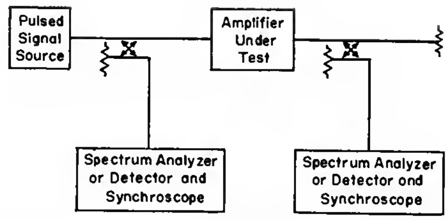


Fig. 56—Arrangement for determination or comparison of pulse shape and spectral distribution.

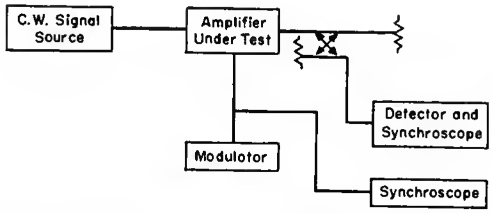


Fig. 57—Arrangement for comparison of output pulse with modulator pulse.

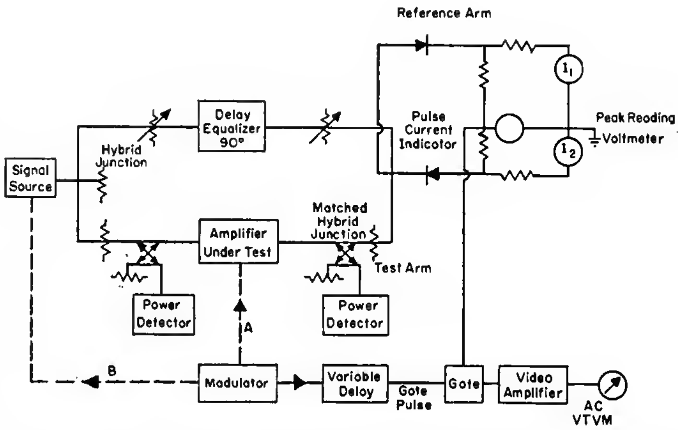


Fig. 58—Arrangement for measurement of amplifier phase jitter. A—modulator driving amplifier for category 1. B—modulator driving source for category 2. A and B—suitably phased for category 3.

The phase bridge shown in Fig. 58 is essentially the same as that shown in Fig. 48 and discussed in Section 23.2, but the circuit is adapted to pulse operation. The same operating procedures and precautions are followed. With pulsed RF input (categories 2 and 3), the input signal is modulated, and the test and reference arms must be carefully adjusted so that their electrical lengths differ by a quarter wavelength. If there is considerable difference in arm lengths, the phase bridge will act as a frequency discriminator, and any frequency jitter in signal source will produce significant changes in video output. A gate circuit is included before the video amplifier to limit the measurement interval to a selected portion of the pulse interval. A trigger signal for the gate is obtained through a variable delay from a pulse modulator that drives either the amplifier under test or the pulsed signal source. The video amplifier must have a bandwidth adequate to accommodate the gating signal without producing excessive transients. The output of the video amplifier can be measured by means of an

oscilloscope, an ac voltmeter or a harmonic analyzer. By the procedure described in Section 23.2, the video amplifier output voltage is converted into degrees phase shift.

29.2.2 Envelope Jitter: Pulse envelope jitter is the variation in the time incidence of the leading edges of the envelopes of successive pulses or in the length of the pulse envelope of successive pulses. Envelope jitter is measured by comparing amplifier output pulses with input RF pulses or electrode-voltage pulses and observing the variation in time delay and pulse duration by the use of conventional pulse techniques.

29.3 Cutoff Characteristics

The attenuation and noise output during the cutoff interval of an amplifier with pulsed electrode voltages is measured using the equipment shown in Fig. 59. In some amplifiers (especially high-power amplifiers) the noise during the cutoff interval is a result of after effects of the main pulse. For this reason, the amplifier must be operated in its normal pulsed condition. In this test the noise output of the amplifier during the cutoff interval is measured by comparing it with the noise output of a standard noise source. A filter is placed at the receiver to limit the frequency range to the band of interest. The filter may be omitted where the receiver itself performs the band-limiting function. The noise output of the amplifier under test may vary with time in the cutoff interval; this may be checked visually by observing the noise output from the receiver by means of an oscilloscope in order to measure the noise level as a function of time. In cases in which the noise output is not affected by the presence of the main signal, the test may be simplified by removing the RF drive and the duplexer and connecting the amplifier output to point *A* of Fig. 59.

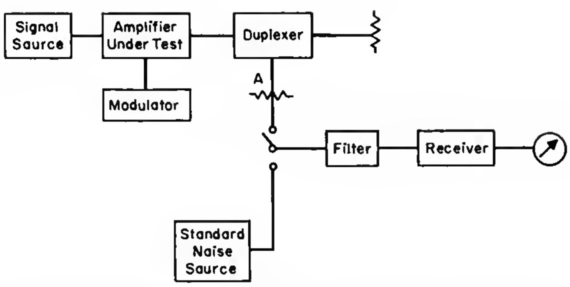


Fig. 59—Arrangement for the measurement of attenuation and noise output of an amplifier during cutoff intervals.

Because the cutoff attenuation, as well as the noise, may be affected by the presence of the main pulse, the measurement of attenuation is also made by means of the equipment of Fig. 59, but input drive is now applied. Attenuation during the cutoff interval is measured by comparing the output of the receiver with the input power. In general, the cutoff attenuation may be a function of the input level used both during the main pulse and during the beam cutoff interval.

29.4 Pulse Echoes (Internal Reflections)

Internal reflections in microwave amplifiers may produce echo pulses. When the echo falls within the main pulse, the effect is to distort the pulse shape. This distortion is measured by use of the procedure of Section 29.1. When the RF input is pulsed and the beam is not pulsed (category 2) and the transit time of the echo through the amplifier is greater than the main pulse length, the echo pulses appear after termination of the main pulse. In this case the quantitites of interest are usually the amplitudes of the echo pulses relative to that of the main pulse, and the time separation between pulses.

Echo pulses are measured by use of the equipment shown in Fig. 60. An RF signal from a pulsed signal generator is sent through the amplifier under test, which is terminated by a matched load. A sample of the output signal is detected and displayed on a synchroscope triggered by the pulsed signal generator. The relative amplitudes and time positions of the pulses can be determined from the synchroscope display.

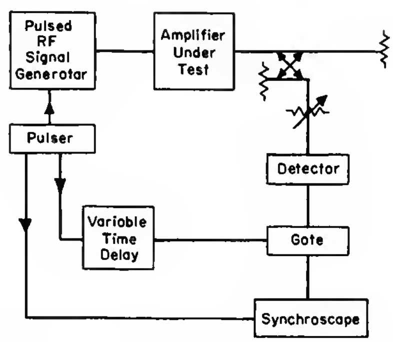


Fig. 60—Arrangement for measurement of echo pulses.

Precautions:

1) In some cases the amplitudes of the echoes may be so small compared with that of the main pulse that simultaneous display on the synchroscope is not feasible. A variable calibrated attenuator and a gate controlled by the pulse source should in such cases be included in the output line. The gate is first adjusted so that the main pulse appears on the synchroscope, and the attenuator is set so that a convenient display of the main pulse appears. The delay of the gate-trigger pulse is then increased by a known amount so that the main RF output pulse no longer appears on the synchroscope; the attenuation is decreased until the echo pulses appear.

2) Because reflections from passive components may also cause pulse echoes, it is important to differentiate between echoes caused by the tube alone, echoes involving reflections between external components, and echoes involving reflections between the amplifier and external components. In measuring echoes caused by the tube alone it is important that all components of the input and output lines be well matched. In specific applications it may also be desirable to measure echoes

when the amplifier operates with known mismatches at its input and output terminals.

30. Test for Voltage-Tunable Amplifiers

All the tests so far described for amplifiers (16–29) are applicable to voltage-tunable amplifiers. However, for these amplifiers many of these tests are conducted with electrode voltages as parameters. Measurements of particular interest for voltage-tunable amplifiers are:

30.1 The Tuning Characteristic

This characteristic gives the frequency of maximum gain as a function of electrode voltages. It is obtained by using the method of Section 16.

30.2 Start-Oscillation Characteristic

This is a characteristic of particular importance for backward-wave amplifiers, which oscillate for beam currents exceeding a critical value, called the *start-oscillation current.* This current is in general a function of beam voltage, output match, and other operating parameters. To obtain adequate gain, backward-wave amplifiers must operate close to start-oscillation current. In order to determine the maximum operating currents of backward-wave amplifiers, the start-oscillation condition should therefore be investigated over a wide range of parameters. This condition is determined by detecting the output of the amplifier under test in the absence of RF driving power. With other operating conditions held constant, the beam current is increased to the value of the threshold of RF oscillations. This value of beam current is the start-oscillation current.

31. Tests for Frequency Multipliers

The test methods described so far (16–29) can be applied to frequency multipliers by modifying the test equipment so that the input and output signals are generated and measured at the proper frequencies. For frequency converters with two inputs (mixers), the tests used for passive mixers are applicable, but in some cases they may have to be modified for the measurement of conversion gain instead of conversion loss.

32. Stability of Characteristics

By *stability of characteristics* is meant the degree to which the tube characteristics remain constant with time. Stability is measured by observing changes of characteristics over a period of time. Generally speaking, stability properties can be divided into three categories. These are:

Long-Term Stability: This characteristic refers to changes taking place in intervals varying from an hour to many thousands of hours. These changes can be caused by such things as accumulation of gas within the tube, changes in cathode activity and evaporation of material from grids. All tube characteristics may be subject to such long-term changes. They are observed by repeated application of the tests already described.

Short-Term Stability: This characteristic refers to changes occurring in intervals varying from a few seconds to one hour. Such effects are usually caused by thermal variations. The variation of any operating parameter of an amplifier, including the input RF power levels, can cause variation of the temperature of the tube elements and this, in turn, can affect tube characteristics (*e.g.*, power output, gain, phase shift through the tube, etc.). The magnitude of these thermal changes and the time required for them to occur are frequently important.

In principle, short-term changes may be measured by tests already described. However, when the time constant of the change is small, care must be taken to insure that measurements are made in a time short compared to this time constant.

Very-Short-Term Stability: This characteristic refers to changes taking place in times typically much shorter than one second. These changes are usually of an electronic nature. In many cases they can be viewed as spurious modulation and measured using the tests of Section 27.

33. Additional Definitions of Terms for Tunable Microwave Oscillators

Tuning Range (of Oscillator). The frequency range of continuous tuning within which the essential characteristics fall within prescribed limits.

Tuning Sensitivity (of Oscillator). The rate of change of frequency with the control parameter (*e.g.*, the position of mechanical tuner, electrical tuning voltage, etc.) at a given operating point.

Tuning Creep (of Oscillator). The change of an essential characteristic as a consequence of repeated cycling of the tuning element.

Response Time (of an Electrically-Tuned Oscillator). The time following a change in the input to the tuning element required for a characteristic to reach a predetermined range of values within which it remains.

Resetability (of Oscillator). The ability of the tuning element to retune the oscillator to the same operating frequency for the same set of input conditions.

Hysteresis. The difference in a characteristic when a tuner position, or input to the tuning element, is approached from opposite directions.

Backlash. The amount of motion of the tuner control mechanism (in a mechanically-tuned oscillator) that produces no frequency change upon reversal of the motion.

Electrically-Tuned Oscillator. An oscillator whose frequency is determined by the value of a voltage, current or power. Electrical tuning includes electronic tuning, electrically activated thermal tuning, electromechanical tuning, and tuning methods in which the properties of the medium in a resonant cavity are changed by an external electrical means. An example is the tuning of a ferrite-filled cavity by changing an external magnetic field.

Part 7

Cathode-Interface Impedance

Subcommittee 7.6

Physical Electronics

H. B. FROST, *Chairman* (1959–1962)
R. M. MATHESON, *Past Chairman* (1955–1958)

R. W. Atkinson
J. G. Buck
L. Cronin
P. N. Hambleton
J. M. Lafferty
J. E. White

1. INTRODUCTION

1.1 General Comments

Although one or more may be negligible under specific conditions, three impedances always exist between the surface of an electron-tube oxide cathode and its external terminal: the cathode-lead impedance, the cathode-interface impedance, and the cathode-coating impedance. Of these three impedances, the cathode-lead impedance is that of an inductance which is fixed by the mechanical design of the tube, but the cathode-interface impedance and the cathode-coating impedance are determined largely by the chemical state of the cathode and consequently change as a tube ages. The magnitude of the coating impedance is relatively fixed from low audio frequencies up to UHF; at higher frequencies the capacitance of the coating shunts the coating resistance. The cathode-interface impedance, however, changes from essentially a pure resistance at 10 kc to essentially a pure capacitance at 10 Mc, and by feedback action this variation is translated into an increase in transconductance as the frequency increases over this range. This variation of transconductance in triodes and pentodes both makes cathode-interface impedance technically important and provides means for its measurement.

Unfortunately from the measurement point of view, the cathode-interface impedance is not provided with an independent set of terminals. Consequently this impedance must be inferred from measurements of other tube parameters. For triodes and pentodes, the frequency dependence of transconductance can be used; for diodes, the frequency dependence of the plate resistance can be used. Over the frequency band in which cathode-interface impedance causes the transconductance to be frequency-sensitive, other effects also cause the transconductance to change with frequency. The phase of the transconductance—more properly, the transadmittance—is dependent upon transit time, cathode-lead inductance, and grid-plate capacitance, as well as the frequency. These effects must all be considered if the cathode-interface impedance is to be measured accurately.

Cathode-interface impedance varies with cathode temperature, current density and distribution, and time; *i.e.*, the interface impedance measured at a given time is affected not only by the conditions of measurement but also by the entire history of the tube prior to the measurement. Hence, it is essential that measurements of interface impedance be made under conditions which closely approximate the actual operating conditions and that the time required for measurement not be unduly protracted.

Considered exactly, cathode-interface impedance is actually a distributed network. However, for engineering purposes the interface impedance may be approximated by either of the lumped-constant equivalents shown in Fig. 1 a and b, respectively. Frequently the

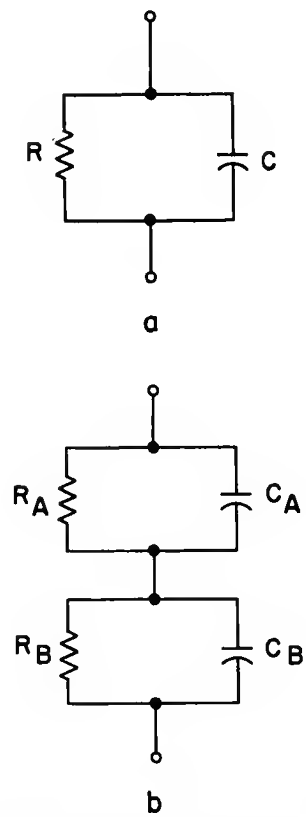


Fig. 1—Networks equivalent to interface impedance

simpler circuit (Fig. 1a) is a satisfactory approximation. All normal methods of measurement are based on the fact that, at a sufficiently high frequency, the impedance of these circuits approaches zero. In practical cathodes having an area of about 1 cm^2, the resistance (R) ranges from less than one ohm to several thousand ohms and the time constant (RC) ranges from 2×10^{-8} to 5×10^{-6} second. The larger resistances tend to be associated with the longer time constants. Ordinarily, the impedance approaches zero at frequencies above 30 Mc, and it approaches R at frequencies below 10 kc. However, at lower frequencies the total cathode impedance may continue to rise because of the frequency dependence of the *cathode-coating impedance*. In CW methods of measuring cathode-interface impedance, frequencies of 10 Mc and 5 kc are frequently used. For transient-type measurements, one can use either a square wave having a period of 2×10^{-5}-second or a rectangular pulse of 1×10^{-5}-second duration. The rise time and fall time in either case should be 3×10^{-8} second or less.

1.2 General Test Conditions

Early in life tubes usually have small or negligible values of cathode-interface impedance. As tubes age, especially under conditions of low plate current, interface impedance may increase, depending upon cathode materials and processing. However, the interface impedance frequently is radically affected by a change of operating conditions. Consequently, in order that correct data may be obtained, the following precautions should be observed.

1) After an aging period, tubes should not be subjected to any other test prior to the interface measurement, unless such test is conducted under actual operating conditions.
2) During measurement, the tube should be operated at voltages and currents as nearly identical with those used in actual operation as practicable. If different conditions must be used, the measuring current should be less than actual operating current, if possible.
3) Cathode temperature must be accurately controlled or determined. A 1 per cent deviation in heater voltage may lead to a 10 per cent error on the measured interface impedance. The cathode must reach thermal equilibrium prior to measurement.
4) Since the measurements are performed in wideband HF circuits, all the normal precautions essential to the use of such circuitry must be observed. In particular, it is essential to eliminate, or correct for, the effects of stray capacitance and inductance that may otherwise invalidate the measurement.

1.3 Measurement Circuits

Five circuits for measuring interface impedance are described in the following paragraphs. These circuits can be divided into three groups: 1) bridge circuits, Sections 1.4, 1.5; 2) pulse-comparison circuits, Sections 1.6 and 1.7; and 3) the CW circuit, Section 1.8. The choice of a measurement circuit must depend on the requirements of a particular application. The bridge circuits described in Sections 1.4 and 1.5, while relatively complex, provide a maximum of information, determining the values of all four elements of the equivalent network of Fig. 1 b, as well as the g_m of the tube. The shunt-admittance method is particularly useful for tubes having very high g_m. The pulse-comparison circuits, described in Sections 1.6 and 1.7, are simpler than the bridge circuits but provide less information about the interface impedance. The differential-comparison method is convenient if a variety of tube types are to be measured. The standard-tube-comparison method is particularly suitable for measurements of large numbers of tubes of the same type. Interface impedance having abnormal time constants can be detected but not measured by these circuits. An abnormal decay curve in the differential-comparison method or a poor null in the standard-tube-comparison method indicates the existence of such unusual impedances. The CW method requires a minimum of equipment and is simple to operate, but provides information about only the resistive component of the interface impedance and can be inaccurate with impedances having abnormal time constants.

The various methods differ somewhat in their ability to measure interface impedances with low resistance—less than 25 ohms—and time constants less than 0.1 μsec. Only the bridge circuits of Sections 1.4 and 1.5 are known to provide accurate data, within about 1 ohm, for impedances of this magnitude.

The areas of application of the various methods are shown in Table I.

TABLE I

Application	Method		
	Bridge	Pulse Comparison	CW
Highest Accuracy, Especially Low Impedance, Short Time Constant	Preferred		
Routine (Production Life Tests)	Satisfactory	Preferred	Satisfactory
Field	Not Portable		Preferred

All interface-impedance measurement methods described are small-signal methods. These have been chosen because they minimize changes of interface impedance caused by the measurement. The amplitude of the signal required to drive these circuits has not been tightly specified because this varies widely with tube type. The signal amplitude used in a measurement

should be stated. Large-signal methods have been used and are useful for specialized purposes. However, large-signal methods may alter the interface impedance of the tube under test unless the test signal is substantially identical with that employed in the actual operation of the tube, and they cannot be recommended for general application.

In general, equipments must be calibrated and tested with dummy networks, since tubes with interface impedance are not stable for many measurements over a period of time. Down to 25 ohms, ordinary precautions are adequate, but special networks with added resistance should be used below 25 ohms [3].

The circuit diagrams (Figs. 2 to 6) show nominal values for the circuit elements. For most of the receiving and small transmitting tubes that employ unipotential oxide-coated cathodes, the nominal values specified will be found satisfactory. Some transmitting tubes, however, will require modifications of load resistances and bias networks in order to allow proper operation.

1.4 Complementary-Network Bridge

The tube is tested in its normal amplifier connection in the circuit of Fig. 2. The grid is driven by a square wave from one-half of a phase splitter. The phase-splitter tube should be a high-transconductance type, such as a WE 417A or a parallel-connected 6J6. The input square wave should have a rise time of 3×10^{-8} second or less and a period of about 2×10^{-5} second. The amplitude should be as small as is consistent with accuracy. The g_m-balance resistor is adjusted to equal the reciprocal of the (low-frequency) transconductance of the tube under test, and the quadrature error is balanced by C_Q while the oscilloscope is observed. If the tube has no cathode interface impedance, there is then no error output signal when all elements of the complementary network are set to zero. If there is interface impedance, the apparent transconductance of the tube is frequency-sensitive and not matched by the reciprocal-g_m resistor, and an output signal appears on the oscilloscope. The unbalanced waveform for a low gain on the oscilloscope is shown in Fig. 2 as insert A. The transconductance is adjusted to close the error prior to the switching point near the left-hand edge of the insert. (The overlap is obtained by causing the oscilloscope to synchronize on an odd subharmonic of the 100-kc synchronizing signal). The quadrature control C_Q is adjusted to remove the spike on the leading edge at the switching point, and the resistance R_1 is adjusted to remove any step at the leading edge, with a finite setting for L_2, the branch R_3-L_3 being open. If the error cannot be reduced to zero by adjustment of L_2, with slight readjustment of C_Q and R_1, an error pattern similar to Insert B of Fig. 2 can be obtained. This error pattern indicates that the interface impedance has two time constants, equivalent to the circuit of Fig. 1b. A balance can be obtained through use of R_3 and L_3, with slight adjustments for R_1 and L_2. When a null is obtained the parameters of the complementary network are directly related to the interface parameters, and the latter can be obtained by calculation [3].

The complementary network shown in Fig. 2 has four adjustable elements and two unavoidable parasitic elements. If the parasitic elements R_2 and L_1 can be neglected, the interface resistance is simply the value of R_1. In general, however, the parasitic elements may not be neglected, particularly for low values of resistance and short (less than 0.1-μsec) time constants. In such cases, the cathode interface resistance can be calculated from the complementary network parameters as follows:

$$R_i = \frac{(L_1L_2)^2R_3 + (L_2L_3)^2R_1 + (L_1L_3)^2R_2}{(L_1L_2 + L_1L_3 + L_2L_3)^2} - \frac{R_1R_2R_3}{R_1R_2 + R_1R_3 + R_2R_3}.$$

If an uncertainty in the value of the stray inductance L_1 is troublesome, the stray resistance R_2 can be padded out with a fixed resistance of 5 or 10 ohms to reduce the effects of uncertainty in L_1. The inductance L_1 should in any case be as small as possible because this inductance is effectively in series with the cathode lead of the tube under test and can, if too large, cause a phase shift so large that it cannot be properly corrected by the quadrature capacitance.

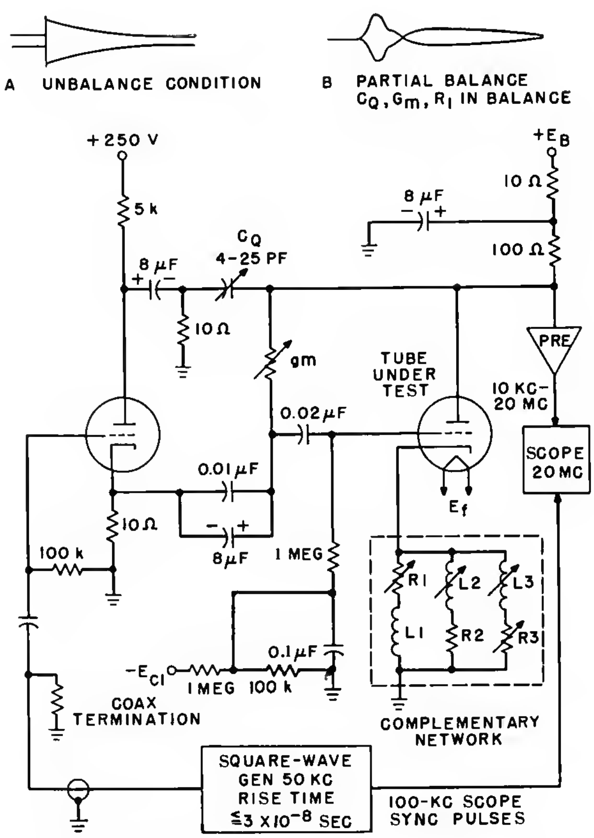
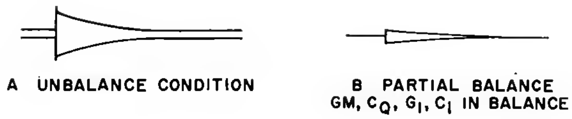


Fig. 2—Circuit arrangement for complementary-network-bridge method.



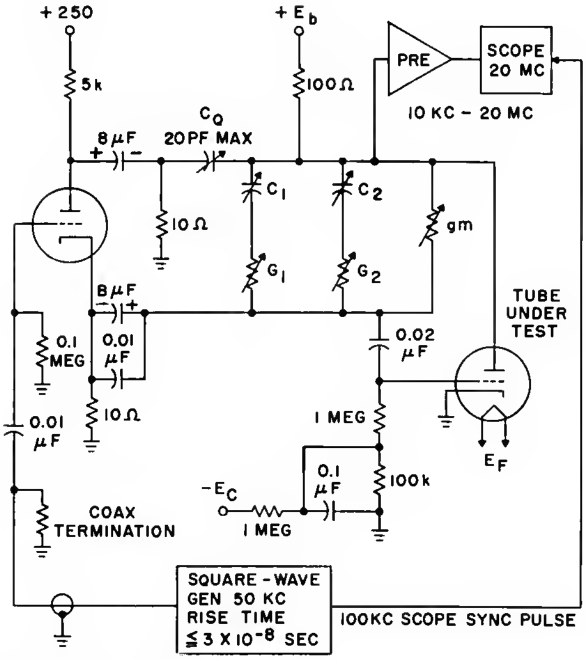


Fig. 3—Circuit arrangement for shunt-admittance-bridge method.

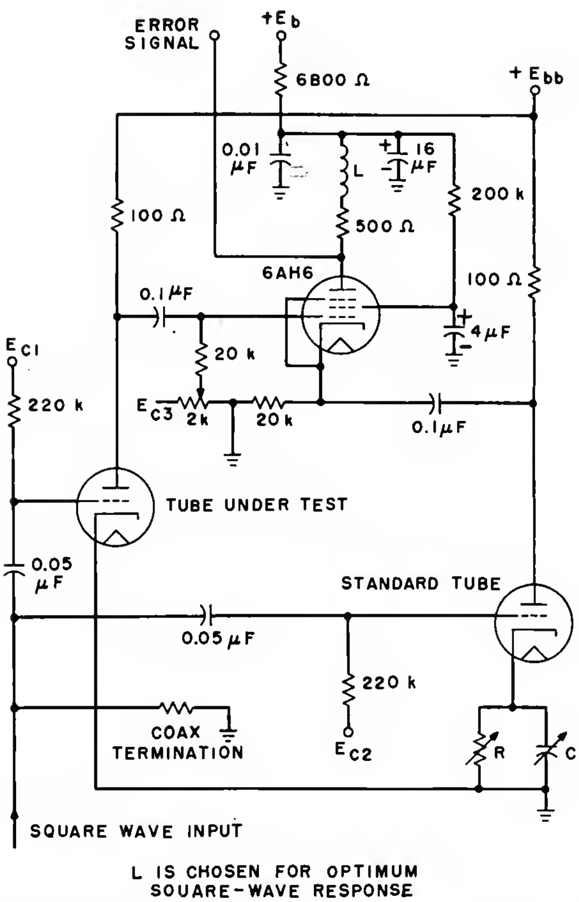


Fig. 4—Circuit arrangement for standard-tube-comparison method.

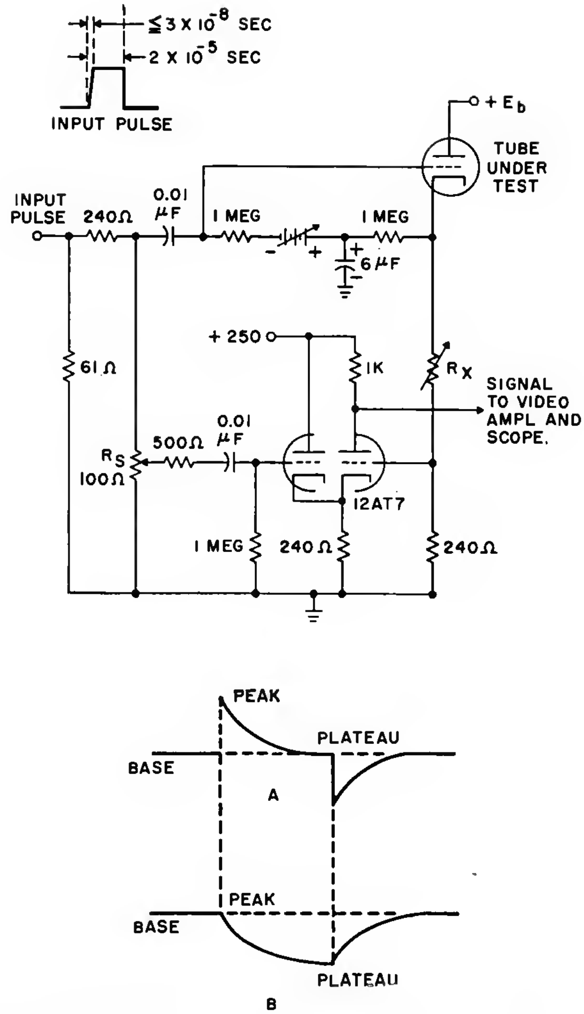


Fig. 5—Circuit arrangement for differential-comparison method.

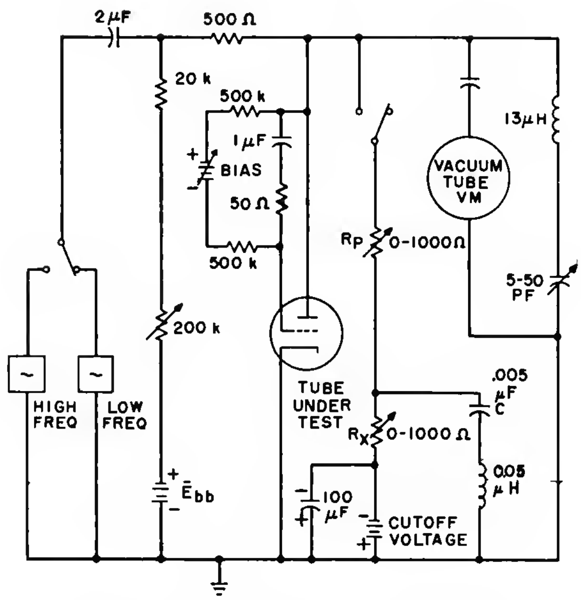


Fig. 6—Circuit arrangement for CW method.

This bridge circuit is capable of achieving great accuracy and providing detailed information about the interface impedance. However, four and sometimes six elements have to be adjusted to achieve a balance.

1.5 Shunt-Admittance Method

The shunt-admittance method for the measurement of interface impedance employs a wideband transconductance bridge similar to that used in the complementary network method. However, instead of a complementary network in the cathode circuit to cancel the variation of transconductance with frequency, the shunt-admittance method actually measures the complex transadmittance in terms of a network that is equivalent to the reciprocal transconductance in series with the interface impedance. Network-impedance levels make this method particularly valuable when tubes with transconductances in excess of 10,000 μmhos are measured, since it is not necessary to simulate or complement the low values of cathode-interface impedance which are important in these tubes. For low values of interface impedance, *i.e.*, R_i about 1 ohm, construction of either simulation networks or complementary networks becomes difficult because stray inductances are very troublesome at low impedance levels.

The tube under test is triode-connected and operated in the circuit of Fig. 3. The grid drive is supplied by one output of a phase-splitter that is driven by a square wave of 2×10^{-5}-second period and 3×10^{-8}-second rise time. The square-wave amplitude should be as small as consistent with good accuracy, about 100 to 200 mv. In balancing, the trailing edges of the square wave, as shown on Insert A of Fig. 3, are brought into coincidence (with the oscilloscope synchronized at an odd subharmonic of the 100-kc synchronizing pulses). With the admittance arms open, the residual capacitance feedthrough is removed by means of the quadrature control C_Q. Then any remaining error signal is caused by cathode-interface impedance and is balanced out by introducing admittance in shunt with the g_m arm. Each of the admittance arms removes one exponential error term that appears after the switching point of the square wave. Insert B of Fig. 3 shows a waveform with one term remaining.

The interface resistance can easily be calculated from the relation

$$R_i = \frac{\mu}{(\mu+1)} \frac{G_1+G_2}{g_m(g_m+G_1+G_2)},$$

where g_m is the LF transconductance and G_1 and G_2 are the HF conductances of the two shunt transadmittance arms. If more detailed information on the interface impedance is required, the interface impedance can be calculated [3] by standard network-synthesis methods from the transconductance and admittances as measured on the bridge.

1.6 Standard-Tube-Comparison Method

The tube is tested as an amplifier in the circuit of Fig. 4. The input square wave should have a rise time 3×10^{-8} second or less and a period of about 2×10^{-5} second. The amplitude should be as small as is consistent with accuracy. The output pulse from the tube under test is compared, by a difference amplifier, with the output of standard tube (preferably of the same type as the tube under test) known to have substantially zero interface resistance.[1] The error output from the difference amplifier is adjusted to a null by varying the bias of the standard tube and elements R and C. When a null is obtained, the values of R and C are equal to those of the interface impedance, within the limits of a two-element approximation.

For interface time-constants greater than 0.1 μsec this method is capable of achieving approximately the same accuracy in the measurement of interface resistance as the complementary-network bridge. Selection and maintenance of standard tubes must be done with care.

1.7 Differential-Comparison Method

The tube is tested as a cathode follower in the circuit of Fig. 5. The input pulse should have a rise time of 2×10^{-8} second or less and a duration of about 1×10^{-5} second. The amplitude should be as small as is consistent with accuracy. The output signal from the tube under test is compared with a fraction of the input signal by means of a differential amplifier. With R_x adjusted to zero, the condition shown in Fig. 5A is obtained by adjusting R_s. The output signal is adjusted to the condition shown in Fig. 5B by adjusting R_x. Then R_x is equal to the interface resistance. The interface capacity may be estimated from the decay curve observed on the scope.

This method is capable of accuracy similar to the preceding method (1.6).

1.8 CW Method

The tube is tested as an equivalent diode in the circuit of Fig. 6. The tube is compared at two frequencies with the analog network R_p, R_x and C. Because only two frequencies are used, only two parameters of the unknown can be determined. Accordingly, the capacitor of the analog network is chosen so that its reactance is small relative to the interface resistance at 10 Mc and large at 5 kc. The residual reactance at 10 Mc is cancelled by a small series inductance. To make the measurement, the tube is first driven by a 10-Mc constant-impedance generator, and the ac voltage across the

[1] A tube having substantially zero interface impedance may be selected by measuring interface impedance with the heater voltage approximately 20 per cent above and below normal. If no significant change of interface impedance is noted, the tube may be considered to have negligible interface impedance. It is essential to recheck the standard tube periodically.

tubes is set to a reference value (about 0.1 volt) as measured with a vacuum-tube voltmeter. The 5-kc generator is then switched in and adjusted to give the same level. With the tube biased off, the analog network is then driven by the 10-Mc source and R_p adjusted to give the same voltage. The same general procedure is repeated with the 5-kc generator, but R_x is adjusted. R_x is the interface resistance of the tube. The bias network, used to obtain suitable current levels and division with triodes and multigrid tubes, is not needed for diodes.

This method generally is capable of determining interface resistance with an accuracy comparable to that of the circuits of Sections 1.6 and 1.7. However, as it does not provide information about the time constant, the results may be inaccurate (low) when impedances having unusual time constants are measured.

2. Bibliography

[1] American Soc. for Testing Materials, "Tentative Method of Test for Interface Impedance Characteristics of Vacuum Tube Cathodes," ASTM Standard F 300-55T; 1955.

[2] W. V. Shipley, "A method of measuring cathode-interface impedance," 1956 IRE National Convention Record, pt. 3, pp. 64–66.

[3] H. B. Frost, "New methods for the measurement of cathode-interface impedance," IRE Trans. on Electron Devices, vol. ED-6, pp. 315–321; July, 1959.

[4] J. R. Tillman, J. Butterworth, and R. E. Warren, "The dependence of mutual conductance on frequency of aged oxide-cathode valves and its influence on their transient response," *Proc. IEE*, vol. 100, pt. 4, pp. 8–15; October, 1953.

[5] H. M. Wagner, "Pulse and sine-wave techniques for measuring cathode interface," *Proc. 4th Natl. Conf. on Tube Techniques*, pp. 214–224; 1959.

Part 8
Camera Tubes

Subcommittee 7.8

Camera Tubes

B. H. Vine, *Chairman* (1955–1962)

J. A. Hall, M. Rome, D. L. Schaefer, R. R. Linden, C. E. Thayer

1. Introduction

The testing of *camera tubes* requires use of the sciences of Photometry, Optics of Lenses, and Electrical Measurements. Considerable care is required in the design, construction, adjustment and operation of test equipment in order that the results obtained will be characteristic of the camera tube only, and further, that the results will be measurements which can be reproduced on a test equipment of different design. Ordinarily the practical accuracy of these tests is limited to somewhere between 1 and 10 per cent.

The results of any test on a camera tube that utilizes magnetic deflection of focus or alignment are dependent on the design of the components used to produce the magnetic fields. The subject of any test must be considered to be the combination of the tube and these field-producing components. These components must be specified in the description of the test.

These methods of test apply to tubes in which the photosensitivity and the scanning functions are incorporated in a single tube. The tubes used in flying spot systems are covered in Parts 2 and 5.

1.1 Definitions

Television Line Number. The ratio of the raster height to the half period of a periodic test pattern.

Note: Both quantities are measured at the camera-tube sensitive surface.

Example: In a test pattern composed of alternate equal-width black and white bars, the television line number is the ratio of the raster height to the width of each bar.

2. Test Equipment

Camera tubes require horizontal and vertical deflection circuits and several adjustable sources of dc voltages. The most widely used camera tubes are those in which a stored electrical-charge image is erased by scanning. This type of tube requires blanking signals to prevent erasure of stored information during the retrace times.

An illuminated test chart[1] and a lens to project the test pattern on the sensitive surface of the tube are required. A calibrated source of illumination is required for sensitivity measurements. A monochromator and calibrated radiation measuring thermocouple are required for the determination of spectral-sensitivity characteristic.

A *line-selector oscilloscope* is used for amplitude response and transfer-characteristic measurements and is convenient for others. This instrument displays the signal from any selected horizontal line, and in addition produces a marker pulse which may be added to the video signal to indicate on the picture monitor tube which line is being displayed on the oscilloscope. A dc oscilloscope or a dc pen recorder is used in the vidicon lag test. A calibrated "bar generator" is needed to measure the signal-output-current amplitude of the iconscope or orthicon. A calibrated dc meter capable of measuring in the nanoampere to microampere regions is required for the direct methods of measurement of photosensitivity.

The performance required of the above equipment will be described with the individual methods of test.

A block diagram of the basic circuit for camera tube tests is shown in Fig. 1.

2.1 Design and Adjustment of Test Equipment

Critical attention must be given to the uniformity of illumination of the test pattern, the quality of the lens used for imaging the test pattern on the tube face, the frequency dependence of amplification and phase shift in signal amplifiers, and the linearity of amplifiers and oscilloscopes.

2.1.1 The Picture Monitor Scanning Amplitudes for test purposes are set so as not to overscan the face of the

[1] The 18 inch × 24 inch resolution gray-scale and linearity opaque test charts for testing camera tubes are obtainable from Electronic Industries Association, Engineering Office, Room 650, 11 West 42 St., New York 36, N. Y. The gray scales on the charts are available with either logarithmic or linear reflectance steps.

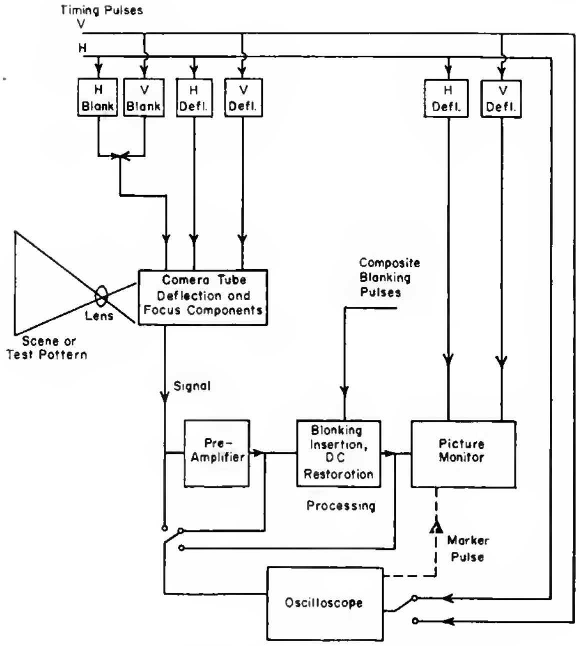


Fig. 1—Block diagram of camera-tube test circuit.

monitor tube in order that the whole raster may be visible at all times.

2.1.2 The Dimensions of the Scanned Area on the sensitive surface of the camera tube (or the equivalent dimensions referred to the sensitive surface in the case of imaging types of camera tubes) may be set to the desired size by the following procedure. The desired scanned area is marked off on a sheet of translucent paper. The lens to be used is removed from the camera and is used to image a rectangular test pattern having the desired aspect ratio on the translucent paper. The lens-to-test pattern distance required to make the image of the test pattern fit the marked area on the translucent paper is determined. The lens iris should be fully open for minimum depth of focus during this determination. The lens is replaced on the test camera, care being taken not to change its focal length and the test pattern set at the determined distance. Any point on the lens barrel may be used as the reference point for these measurements, as long as it is used consistently.

With the camera tube in both optical and electrical focus, and with the scan directions rotated to align with the test rectangle, the scan amplitudes are set to scan the test rectangle by watching the picture monitor.

If it is desired to know the dimensions of the scanned area on a camera tube in a camera in operation, the above procedure may be reversed.

The above procedure is proper for setting or determining the scanned area for resolution tests. In the case of tests involving measurement of dc target or photocathode currents, the picture monitor does not display the whole area scanned on the camera tube because blanking insertion trims the edges of the picture. If this error is significant, the camera scan should be set to scan the test pattern rectangle using the output of the preamplifier on an oscilloscope as an indicator of proper scan.

2.1.3 The Linearity of Scan is particularly important when measuring the signal uniformity in storage type camera tubes because the signal is proportional to the rate of scanning. The deflection circuits may be adjusted to give a linear scan on the camera tube by means of the EIA Linearity Chart as described in the "IRE Standards on Television: Methods of Measurement of Aspect Ratio and Geometric Distortion, 1954" (54 IRE 23. S1).[2] Since lens distortion of the barrel or pin-cushion type affects this adjustment, it should be determined that this is not a significant factor.

2.1.4 Determination of the Number of Loops of Focus in the Beam may be made based on the fact that this number is proportional to the magnetic-focus-field strength and inversely proportional to the square root of the potential of the space through which the beam passes. While this statement is exact only for a uniform magnetic focus field and constant potential, it is suitable for use in making this determination in commonly used camera tubes.

With fixed voltages on the tube electrodes, the focus coil current is varied and its value determined for two or more successive conditions of electron beam focus. If the coil currents for these focus points are plotted against their numbers, the straight line so determined may be extrapolated to zero current. This point identifies the zeroth loop of focus, and the ordinal number of all the focus points is then determined. (It is assumed that the magnetic field is proportional to the coil current, which is true except in the unusual case where there is a large amount of hard ferromagnetic material in the flux path.)

Alternatively, observations may be made at fixed focus field while varying the voltage for focus points. If more than one electrode determines the beam-space voltage, the ratios of the voltages on all these electrodes must be kept fixed. The reciprocal of the square root of the voltage is plotted against focus number and extrapolated to zero (infinite voltage) to determine the zeroth loop.

2.2 *Specification of Test Results*

To make test results meaningful, certain parameters and conditions used in making the test are significant and should be specified with the test results. In general, the horizontal and vertical scan rates are significant. If shading corrections or aperture corrections have been added to the signal before evaluation, these should be specified. If pulsed light is used rather than continuous light, the details of the pulse should be described.

[2] Proc. IRE, vol 42, pp. 1098–1103; July, 1954.

For each test described in Section 3, the more significant parameters and conditions are listed. Unless otherwise specified, all electrode voltages are measured with respect to the thermionic cathode.

3. Methods of Test

3.1 Measurement of Transfer Characteristic

3.1.1 Flat-Field Method: The tube face is uniformly illuminated and the signal-output current is measured by means of a dc meter in the camera-tube output. Transfer characteristic is determined by plotting signal-output current as a function of illuminance. This method ordinarily is suitable only for the vidicon.

This method is similar to that used for measurement of luminous sensitivity of phototubes; see Part 5, Section 2 (2.1, and 2.1.2), for the use of standard lamps, filters, calculation of light flux, etc. It will be noted that no lens is used for this test so the lens normally on the test camera must be removed. The scanning must be on during this test with the beam in focus.

A rectangular masking aperature having the dimensions of the normal scanned area is placed directly against the tube face. The vidicon deflection is adjusted so that the illuminated rectangle is just overscanned.

The following information should accompany the results of any determination of vidicon transfer characteristic: spectral distribution of light over the region of significant photosensitivity; signal-electrode voltage; dark current; tube-face temperature and raster size.

3.1.2 Gray-Scale Method: A horizontal gray scale, comprising steps of known reflectance value (or transmittance value if the chart is illuminated by transmitted light), one of which is a suitable black,[3] is uniformly illuminated and imaged on the sensitive surface of the tube. The signal from the horizontal lines passing through the gray scale is observed on an oscilloscope, or a single line on a line-selector oscilloscope. By plotting signal amplitude vs the reflectance of the corresponding step, a transfer characteristic may be obtained which is relative, both in light and in signal. This may be made absolute by suitable calibration techniques, described in 3.1.3 and 3.1.4. It should be determined that errors introduced by amplifier nonlinearity are not significant.

3.1.3 Signal Measurement by Calibration Pulse: A pulse of known current or voltage may be introduced at the coupling resistance (the camera-tube load resistance) to be used for comparison purposes at the oscilloscope so that absolute signal magnitudes may be measured.

3.1.4 Determination of Camera-Tube Illumination: The illuminance on the sensitive surface of the camera tube, in relation to the scene illuminance, can be determined by the following approximate relationship:

$$I_{pc} = \frac{TRI_s}{4f^2(1+m)^2},$$

where

I_{pc} = photocathode illuminance,
f = the ratio of the lens focal length to the lens-iris diameter,
I_s = scene illuminance,
T = total transmission factor of lens,
R = reflectance of the step in the gray scale,
m = linear magnification from scene to target.

Scene illuminance I_s may be measured by means of a conventional light meter. Sections 3.1.2, 3.1.3, and 3.1.4 together represent a method for determining the absolute transfer characteristic of any camera tube.

3.1.5 Note on the Effect of Redistribution: In tubes in which redistribution effects occur, such as the image orthicon and iconoscope, the signal from a given picture area is influenced, not only by the illumination on that area, but also by the illumination on surrounding areas. Thus such a tube has no unique transfer characteristic. Frequently no attempt is made to give a detailed transfer characteristic for such a tube. An operational test is made using a gray-scale chart, as in 3.1.2, with the gray scale imaged to occupy only a small fraction of the raster area. The remainder of the raster is a uniform gray background.

For the image orthicon, the illuminance at the knee is sometimes specified, rather than a complete transfer characteristic.

3.1.6 Knee of the Image-Orthicon Transfer Characteristic: The oscilloscope display of a gray scale is observed while the illuminance on the tube is increased by opening the lens iris, or by other means. The highest reflectance step in the chart is considered to have reached the knee when its signal no longer increases at the same rate as those of lower reflectance steps. In many cases, the knee can be defined within one-half lens stop (*i.e.*, within a factor of 1.4).

The following information should accompany the results of any test on image orthicon transfer characteristics or knee: target voltage above target cutoff; the size of the image of the gray scale on the tube sensitive surface relative to the raster size and the reflectance of the background surrounding the gray scale; and if absolute light values are quoted, the spectral distribution of the light used.

3.2 Measurement of Noise

It is usually necessary to measure the noise at the output of a camera tube in the presence of large signals which are not to be considered as noise, such as may be present during blanking intervals or due to spatially stationary signal variations which are to be measured separately.

A single horizontal line of the camera signal is displayed on the line-selector oscilloscope. A portion of this

[3] This may be a swatch of clean black velvet, a blackbody cavity or an opaque strip across the tube face, so that the light reaching the sensitive surface is not significant.

line having uniform signal selected for measurement is imaged on a slit parallel to the sweep direction. The light passing through this slit is measured by means of a photocell whose output is read on a meter. An arrangement, either mechanical or electrical, is used to produce known displacements of the slit relative to the image of the trace.

The light meter reading may be plotted as a function of the displacement. If there is no noise present on the oscilloscope trace, such a curve is a single narrow peak whose width is determined by the oscilloscope tracewidth and the slitwidth. If noise is present on the trace, then this is usually a symmetrical bell-shaped curve. If displacements are called x and measured from the center of this curve, then, at a given displacement, the meter reading called y is proportional to the fraction of the time the output of the camera falls between x and $x+\Delta x$ away from its mean value, where Δx is determined by the tracewidth and the slitwidth. The rms value of x may be determined from the definition of rms value:

$$\sqrt{\overline{x^2}} \equiv \sqrt{\int_{-\infty}^{\infty} x^2 y dx \Big/ \int_{-\infty}^{\infty} y dx}.$$

The integrals may be determined graphically. However, simpler methods are possible if the curve is closely Gaussian; that is, y is proportional to $\exp -\frac{1}{2}(x/\sigma)^2$, where σ is a constant which determines the width of the distribution and is adjusted to obtain a fit. It is readily shown that for this distribution the rms value is equal to σ. Where it is known that the distribution is Gaussian, then the rms value may be determined from a few points; for example, the two points at which $y=0.6065$ of its peak value are a displacement 2σ apart.

The slit and tracewidths should be small compared to the rms value to be measured. The effects of saturation of phosphor, halation, light scattering and stray light, and nonlinearity of response of photocell should be made insignificant. If statistical variations make the meter unsteady, the time constant of the photocell-meter system may be lengthened. A calibration pulse introduced at the camera-tube output, as in 3.1.3, allows values of x to be converted into terms of output current. Often only the SNR is desired. In this case, a black-to-white transition in the test pattern is used as the reference magnitude. Since the noise may be much smaller in magnitude than the signal, each may be measured against an adjustable calibration pulse in the oscilloscope.

It should be noted that the usual effect of the noise on an observer is a function, not only of the signal-to-rms noise ratio, but also of the shape of the noise power-vs-frequency spectrum. Tubes yielding similar noise spectra may be compared on the basis of their signal-to-rms noise ratios.

3.2.1 Measurement of Image-Orthicon Noise: The test of 3.2 may be used. The noise present in the amplifier should be determined by making a measurement with the beam biased off. If amplifier noise is significant, a suitable correction should be made.

The following information should accompany the results of any noise or signal-to-noise measurement made on the image orthicon: whether beam current in excess of that required to just discharge test-chart whites was used; illuminance in the test whites relative to the illuminance of the knee of the transfer characteristic; the response-vs-frequency characteristic of the test equipment; as modified by the impedance of the camera tube, the scanning frequencies and dimension of the scanned area relative to the target diameter.

3.2.2 Note on Vidicon Noise: The dominant noise in a vidicon-amplifier system is that due to the first few stages of the amplifier. The noise of the vidicon is not usually measured for this reason.

3.3 Measurement of Resolution

3.3.1 State of the Art: It would be very desirable to have a method of arriving at a single number which would specify the resolution of a given imaging process. Numerous methods have been proposed. Two methods[4,5] for obtaining a single number for any given imaging process have been described. Unfortunately, these methods require the use of a sine-wave test chart which is difficult to construct[6] and not generally available. The following tests are only those based on the use of available square-wave charts and arrive at a curve of square-wave amplitude response vs television line number.

The measurement of amplitude response in a camera tube is preferably made in a system all of whose other components together are sufficiently beyond the capabilities of the tube so that they do not significantly affect the results. By adequate design and careful adjustment, the electrical circuits may be made to do this. However, it is often difficult to obtain a lens which will not significantly degrade the picture. In such a case, it is always possible to obtain a lens which will not cover the whole useful image area of the camera tube, but which will have adequate resolution in the area it does cover. It is then necessary to move the lens in order to make tests in various parts of the tube sensitive area. The lens used for such a test is ordinarily a high-quality low-power microscope objective. Camera tubes often have a spectral response which may be broader than the range over which the lens is corrected for chromatic aberration.

3.3.2 Measurement of the Horizontal Square-Wave Amplitude-Response Characteristic: An EIA (Electronic

[4] O. H. Schade, "Image analysis in photographic and television systems," *J. SMPTE*, vol. 64, pp. 593–617, particularly pp. 602–605; November, 1955.

[5] W. N. Sproson, "New Equipment and Methods for the Evaluation of the Performance of Lenses for Television," BBC Monograph No. 15, pp. 11–16; 1957.

[6] N. S. Kapany, *et al.*, "Production of sinusoidal test charts," *J. Opt. Soc. Am.*, vol. 47, p. 103; January, 1957.

Industries Association) test chart having resolution wedges is used. The chart-to-lens distance is adjusted so that the image of the test rectangle just fits the standard raster area on the sensitive surface of the tube by the method in paragraph 2.1.2, or other suitable means. The output of the camera tube is displayed on a line-selector oscilloscope which is adjusted to display the signal from a horizontal line passing through one of the resolution wedges having vertical bars. A marker pulse is convenient for determining what part of the wedge is displayed. The signal amplitude is observed on the face of the oscilloscope. The camera or test chart is then moved so that a transition between a large area white and a large area black appears on the oscilloscope. The ratio of the wedge signal amplitude to the black-to-white signal amplitude is recorded. Since neither the chart illumination nor the camera-tube sensitivity may be very uniform, the best way to move the test area for this test is to move only the test chart.

Often, an arbitrary method is used to select a single television line number to be assigned to the whole response characteristic. For example, the television line number corresponding to the square-wave amplitude response of 0.5 may be selected and called half-amplitude resolution. Similarly, a 10 per cent amplitude resolution may be defined. The "limiting resolution" where the lines in a test wedge can just be discerned in the reproduced picture is frequently stated, but is of doubtful significance.

To permit analysis of deflection defocusing and off-axis aberrations, the half-amplitude resolutions are specified as follows:

a) At the center of the raster with best focus at the center.
b) Near the corner of the raster with best focus at the center.
c) Near the corner of the raster with best focus at this point.

3.3.3 Amplitude-Response Characteristic in the Presence of Redistribution: (Compare with paragraph 3.1.5.) Tubes having redistribution will not have a unique amplitude-response characteristic and may show amplitude responses greater than 100 per cent under some conditions. In an effort to get reproducible results with the image orthicon, the following operational method is sometimes adopted. A test wedge or a set of test bars, together with nearby black and white reference areas are placed on a gray background having 50 per cent of the white reflectance. The background is large enough to cover the entire raster, but the test area is only a small fraction of the raster area.

3.3.4 Information to Be Specified with the Results of Amplitude-Response Tests on Image Orthicons:

1) Raster dimensions at the target relative to target size.
2) Ratio of the highlight illuminance on the photocathode to the illuminance at the knee of the transfer characteristic.
3) Any two of the following:
   a) Current in the focus coil or some other specification of the focus field,
   b) Beam-focus voltage,
   c) *Number of loops of focus* in the scan section of the tube.
4) Target voltage above target cutoff.
5) Photocathode voltage or the *number of loops of focus* in the image section.
6) Beam current in excess of that required to just discharge the highlights.
7) Any deviation of focus, decelerator, or persuador voltages from their values for optimum picture conditions.
8) Tube-envelope temperature near the target.

3.3.5 Information to Be Specified with the Results of Amplitude-Response Tests on Vidicons:

1) Raster dimensions.
2) Total highlight signal output plus dark current.
3) Beam current in excess of that required to discharge the highlights.
4) Any two of the following:
   a) Current in the focus coil or other specification of focus field,
   b) Beam-focus voltage,
   c) *Number of loops* of focus of beam.

3.4 Measurement of Persistence Characteristic

3.4.1 Measurement of Lag: Usually the decay of signal after a step change of illuminance from a highlight value to darkness is the characteristic measured. The scanned area or a portion of it is uniformly illuminated; the light is then cut off and the signal-output current on each successive scan is measured relative to its initial value by means of a dc oscilloscope or oscillograph. When the entire scanned area is used, a fast dc pen recorder may be used in place of the oscilloscope.

The device used to cut off the light may be a rotating light chopper or test pattern. This gives a cyclical test. The cycle time should be long compared to the length of time during which the decay is to be measured to allow ample time for reestablishing the initial state of the camera tube before each test cycle.

When the details of the decay in the first few scans after light cutoff are of interest, a partially illuminated scan interval should be avoided by providing some means of phasing so that light is cut off just after the scan beam passes the portion of the raster to be observed.

3.4.2 Information to be Specified with the Results of Lag Tests:

1) Length of time used to establish the initial state of the camera tube prior to making the step change.
2) Initial and final illuminance (expressed as fractions of that at the transfer-characteristic knee for the image orthicon).

3) Initial signal current (not required for the image orthicon).
4) Raster dimensions at the target (relative to target size for the image orthicon).
5) Tube-envelope temperature near the target.

3.4.3 Measurement of Retained Image (Sticking): Tests for this characteristic are not well developed. The most promising type of test is that in which a reference area of known reflectance or transmittance is placed adjacent to the test area on the test chart. The time required after a standard change of the test-area reflectance for the two areas to reach an equal-signal condition, as observed on the picture monitor, gives one point on the signal change vs time curve.

An indication of image retention may be obtained by holding a black and white scene fixed on the sensitive surface for a preassigned time. A uniform white or gray scene is then substituted, and the time required for extinction of the scene on the monitor observed.

The following information should accompany the results of retained image tests: the application time of the initial scene; the illuminance on the face of the camera tube for the initial scene (expressed as a fraction of that at the transfer-characteristic knee for the image orthicon); the temperature of the tube envelope near the target; in the case of the vidicon, the initial signal current.

3.5 Measurement of Spectral-Sensitivity Characteristic

3.5.1 Spectral Sensitivity of the Image Orthicon: The spectral sensitivity of the image orthicon is proportional to that of its photocathode. The spectral sensitivity of the photocathode is determined by operating the image orthicon as a photodiode, using the image-focus electrode and target assembly as the anode.

3.5.2 Spectral Sensitivity of the Vidicon: The vidicon must be operated as a camera tube, that is, with normal electrode voltages and with focus and deflection in order to obtain a reliable signal output. The basic equipment described in Part 5, Section 2.4, is needed for illumination of the sensitive surface with known amounts of monochromatic light. Light-chopping schemes are not as convenient as for phototubes. In most vidicons the transfer characteristic is nonlinear. When this is true, it is important that that area of the sensitive surface under test be uniformly illuminated, and that either the irradiance (radiant power per unit area) or the signal-output current be held constant. The usual method is to adjust the light to obtain a specified constant signal-output current. If a considerable fraction of the scanned area is illuminated, the signal output may be measured directly, as in Section 3.1.1.

The following information should accompany the results of a spectral-response test on the vidicon: the dark current; the raster size; the value of the fixed signal current; spectrum bandwidth or other specification of spectral distribution.

3.6 Miscellaneous Tests

3.6.1 Beam-Control-Grid Cutoff Voltage: This voltage is measured by making the beam-control-grid voltage more negative until the target receives no appreciable beam current. Complete cutoff of a scene containing blacks is a rough test. A more precise method is to overscan the target so that the corners of the picture represent zero target current. The point at which the image of the target area just disappears into the true zero-current area is beam-control-grid cutoff.

In order to measure a true value of beam-control-grid cutoff voltage on a dc meter any blanking on this grid must be removed because the blanking waveform modifies the operating plateau voltage.

> *Note:* The beam-control-grid cutoff voltage measured above is not necessarily the same as the cutoff voltage determined by reducing the cathode current to zero.

3.6.2 Beam-Control-Grid Operating Voltage: The beam-control-grid voltage is increased until the beam just discharges the highlights in the picture. This end point is observed when the details in the highlights appear. Since the beam-control-grid operating voltage depends on target potential, beam alignment, and illuminance in the highlights, these quantities must be specified.

When beam-control-grid blanking is used, it is not desirable to remove blanking in making this measurement. Consequently, only a relative value is obtained.

The correct value of beam-control-grid modulation (beam-control-grid operating voltage minus beam-control-grid cutoff voltage) may be obtained by measuring both with the same blanking.

3.6.3 Target Cutoff Voltage: This voltage is measured by reducing the target potential until the picture just disappears. This cutoff may vary for different parts of the picture.

A true value of target-cutoff voltage is obtained only when any blanking on the target is removed during measurement.

3.6.4 Alignment: The magnitude of the field required to produce beam alignment is characteristic of each individual tube. Alignment is properly stated in terms of the magnitude and direction of the alignment fields. Ordinarily it is stated in terms of the currents in alignment coils.

Alignment is accomplished by adjusting the field strength or geometrical positions of properly oriented magnetic or electrostatic fields. Proper alignment is ascertained by observing some identifiable information in the scene as the beam is focused and defocused. The alignment fields are adjusted until no rotation or displacement of the information near the center of the scene is noted when the focus is varied.

Example: In an image orthicon without a field mesh, the target is cut off, or the photosurface is placed in the dark, and the beam is focused to observe the dynode aperture. The alignment is accomplished by varying the

current through two coils in the vicinity of the aperture which produce fields at 90° to each other and normal to the longitudinal focus field. The electron-beam focus is varied and the image of the aperture is observed. The alignment-coil currents are adjusted until the image of the dynode aperture is neither displaced nor rotates. The square root of the sum of the squares of the magnitudes of the currents in each coil is taken as a measure of the alignment field required to align the tube properly.

3.6.5 Per Cent Beam Modulation in the Image Orthicon: The photocathode is exposed to uniform illumination, but is masked so that light falls only on the area of the photocathode corresponding to the scanned area of the target.

The following information should accompany the results of an image-orthicon beam-modulation test: target voltage above target cutoff; illuminance on the photocathode with respect to the knee of the transfer characteristic; any setting of the decelerator electrode voltage or alignment other than those for optimum picture; whether any beam current in excess of that used to just discharge the target with light on was used.

3.6.6 Target-Assembly Current Amplification in an Image Orthicon without Field Mesh: The current amplification at the target assembly is the ratio of that part of the beam current collected by the target to the photoelectron current incident on the target, for uniform illumination on the photocathode, where the currents are averaged over a time long compared to the frame time. The target-electrode current and the photocathode currents are measured with the photocathode exposed to uniform illumination, but masked so that light falls only on the area of the photocathode corresponding to the scanned area of the target. The part of the beam current collected by the target is not directly measurable, but is the difference between the target-electrode current (with sufficient beam current to discharge the target) and the photocathode current. In view of the small currents involved, leakage currents may be significant. Photocathode current may be measured in the target circuit by biasing the beam control grid beyond cutoff.

The following information should accompany the results of any measurement of target-assembly-current amplification; illuminance on the photocathode (expressed as a fraction of that at the knee of the transfer characteristic); target voltage above cutoff; photocathode voltage.

3.6.7 Multiplier Current Amplification in an Image Orthicon without Field Mesh: With the tube operating normally the target is cut off or the tube shielded from all light to insure that all beam electrons return to the first dynode, and the anode current is then measured. The beam current is then measured by removing the focus field and measuring the total current to the repeller electrode and the beam-focus electrode. The presence of a small steady deflecting field during measurement of the beam current is used to insure that none of the beam electrons is returned to the first dynode.

The following information should accompany the results of any measurement of multiplier current amplification: anode and dynode voltages; any setting of repeller-electrode voltage away from optimum picture condition during anode-current measurement.

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Part 9

Noise in Linear Twoports

Subcommittee 7.9

Noise

H. A. Haus, *Chairman* (1956–1962)

W. R. Atkinson	W. H. Fonger	W. W. McLeod
G. M. Branch	W. A. Harris	E. K. Stodola
W. B. Davenport	S. W. Harrison	T. E. Talpey

1. Introduction

Spurious undesired signals are always present in signaling systems and their components. These spurious signals are usually called noise. Since noise reduces the amount of information that can be transmitted with a specific signal power, quantitative measures of noise are often indispensable to engineering evaluations of signaling systems.

Test measurements for noise might logically be expected to begin with some generally useful quantitative measure of information-handling capacity in the presence of noise, or of the amount of annoyance which it creates. Such tests can be made in specific instances, but numerous difficulties are encountered. Often they are subjective and hard to evaluate accurately. Furthermore, the effects of noise vary enormously, depending on the system in question and on the levels and characteristics of both signal and noise. It is useful, therefore, to measure noise in the engineering terms commonly used to describe signals.

Measures in terms of power are particularly comprehensive for stationary Gaussian noise, which is completely characterized by its power spectral density. *Thermal noise*[1] and *shot noise* under constant operating conditions are examples of stationary Gaussian noise. Any noise that is steady or stationary in character, and which originates from the linear superposition of a large number of small independent events, is almost certainly Gaussian. Other types of noise exist that may not be Gaussian—for example, "impulse" noise from ignition systems, atmospheric noise, powerline hum, and crosstalk. Complete characterization of these types requires, in addition to spectral density, other information such as the waveform of a typical pulse or the phase of the interference. However, for practical purposes, treatment by the methods below is often adequate; if not, special methods, not described here, will be necessary to deal with these types of non-Gaussian noise.

For simplicity, linearity of system response is assumed in the test methods to be described. Since noise usually enters the system at a point where the signal level is low, linearity will generally exist in the sense that the principle of superposition applies; that is, signals and noise produce currents and voltages that are simply additive, without the complicated intermodulation effects that occur in nonlinear systems. When the noise arises from independent uncorrelated sources, the spectral densities of these independent noises are also additive in a linear system. Heterodyne systems, incorporating frequency shifters, are linear in this sense when the signals are small.

2. Noise Factor

The system whose noise performance is under consideration here is essentially the linear twoport transducer. Signals enter at the input *port*, are processed internally, and leave the transducer at its output port. Any noise input accompanying the signal input is processed in an identical manner. Noise sources internal to the transducer contribute an additional noise at the output port that is independent of the signal or noise input. The noise performance of the transducer is commonly rated by comparing the noise-power outputs of the actual transducer and of its noise-free equivalent.[2] One such measure of performance is the *noise factor* (*noise figure*). The noise factor, at a specified input frequency, is defined as the ratio of 1) the total noise power per unit bandwidth at a corresponding output frequency available at the output port when the noise temperature of the input termination is *standard* (290°K) to 2) that portion of 1) engendered at the input frequency by the input termination. The standard noise temperature 290°K approximates the actual noise temperature of most input terminations.

[1] IRE standard definitions of italicized terms may be found in "IRE Standards on Electron Tubes: Definitions of Terms, 1957," 57 IRE 7.S2, Proc. IRE, vol. 45, pp. 983–1010; July, 1957.

[2] The noise-free equivalent of a transducer is a fictitious transducer with identical port-to-port transfer properties, but with no internal noise sources.

2.1 Variation of Noise Factor with Source Admittance

As defined, the noise factor depends upon the internal structure of the transducer and upon its input termination, but not upon its output termination. Thus, the noise performance of a transducer is meaningfully characterized by its noise factor only if the input termination is specified. The noise factor F of any linear transducer, at a given transducer operating point and input frequency, varies with the admittance Y_s of its input termination (called "source admittance" hereafter) in the following manner:[3]

$$F = F_0 + \frac{R_n}{G_s} \mid Y_s - Y_0 \mid^2, \qquad (1)$$

where G_s is the real part of Y_s, and the parameters F_0, Y_0, and R_n characterize the noise properties of the transducer and are independent of its input termination. Thus, the noise performance of a transducer can be meaningfully characterized for all input terminations through specification of the parameters F_0, $Y_0 = G_0 + jB_0$ and R_n.

The "optimum noise factor" F_0, at the given transducer operating point and frequency, is the lowest noise factor that can be obtained through adjustment of the source admittance Y_s. The "optimum source admittance" $Y_0 = G_0 + jB_0$ is that particular value of source admittance Y_s for which this optimum noise factor F_0 is realized. These interpretations of F_0 and Y_0 are evident from (1). The parameter R_n is positive and has the dimensions of a resistance. This parameter appears as the coefficient of the $\mid Y_s - Y_0 \mid^2$ term in the general expression for F and, therefore, characterizes the rapidity with which F increases above F_0 as Y_s departs from Y_0.

The necessity of using more than one parameter to specify the noise properties of linear transducers was not at first widely recognized. In the LF triode, for example, a single noise generator characterized by an equivalent noise resistance describes its noise properties with sufficient precision over the range of circuit parameters in which the tube is generally employed. The necessity for additional parameters did not become acute until transistors and HF electron tubes were developed.

The parameters F_0, Y_0 and R_n can be calculated if the noise theory of the transducer is known, or, alternatively, can be determined empirically from noise measurements. The empirical method is discussed in Section 4. Until this section is reached, it will be assumed 1) that the input termination of the transducer is specified, and 2) that the noise factor under discussion is the noise factor appropriate to this particular input termination. Used in this context, the noise factor meaningfully characterizes the noise performance of the transducer.

2.2 Average Noise Factor

In any communication system the signal is distributed over some finite bandwidth over which both signal and noise time averages may vary with frequency. Theoretically, in treating the interfering effect of the noise, it would be necessary to consider the frequency distributions (spectra) of both noise and signal; but in practice many cases are sufficiently well approximated by considering only the total powers of signal and noise. When only the total noise power in the band need be considered, the noise performance of the transducer is again rated by comparing its actual noise output for some standard noise input to that noise output which would have been obtained had the transducer been noiseless. One such measure of performance is the *average noise factor*. The average noise factor is defined as the ratio of 1) the total noise power delivered into the output termination by the transducer when the noise temperature of the input termination is standard (290°K) at all frequencies to 2) that portion of 1) engendered by the input termination. For heterodyne systems, 2) includes only that portion of the noise from the input termination which appears in the output via the principal-frequency transformation of the system, and does not include spurious contributions such as those from an image-frequency transformation.

The quantitative relation between the average noise factor $\overline{F}$ and the noise factor $F(f)$ is

$$\overline{F} = \frac{\int F(f)G(f)df}{\int G(f)df}, \qquad (2)$$

where f is the input frequency and $G(f)$ is the *transducer gain*. The average noise factor is the weighted average of the noise factor over the band in question, the weighting factor being the transducer gain. To emphasize that the noise factor, as opposed to the average noise factor, is a point function of frequency, the term *spot noise factor* may be used. Either average or spot noise factor is a numeric, designating a power ratio, which may be expressed in decibels by multiplying its common logarithm by 10.

The average noise factor also depends upon the internal structure of the transducer and upon the admittance of its input termination, but not upon its output termination, except insofar as the output power mismatch varies with frequency, and thus modifies the frequency dependence of transducer gain.

The noise-factor concept is useful over the entire frequency range of engineering interest. Since it compares the actual noise with the fundamental limit set by thermal agitation, the noise factor gives a broad and direct evaluation of the degree to which a system approaches the noise-free ideal.

3. Measurement of Average Noise Factor

In order to measure the average noise factor, it is necessary to obtain a measure of the noise power that is

[3] See tutorial paper in Appendix.

actually delivered to the output termination. This measure is divided by a similar measure of the output noise that would have been obtained if the transducer were noiseless and merely transmitted the *thermal noise* of the input termination at standard temperature. The following general methods of measurement are suitable for performing the evaluation. In each method a measure of the output noise is taken directly, but the methods differ in the ways of determining the reference noise output of the ideal noiseless transducer.

a. CW-Signal-Generator Method (*See Section 3.1*): In this method a power meter at the system output and a calibrated signal generator at the system input are used to determine the frequency dependence of transducer gain. From this dependence a "noise bandwidth" is determined. By use of this noise bandwidth, that portion of the output noise resulting from the input-termination noise can be determined. Dividing total output noise by this reference noise gives the average noise factor.

b. Dispersed-Signal-Source Method (*See Section 3.2*): In this method a signal generator having its available power dispersed uniformly over the pass band of the system, and calibrated in terms of available power per unit bandwidth,[4] is used to determine that portion of the output-noise power which results from the input-termination noise. Suitable dispersed-signal generators are thermionic-noise diodes, gas-discharge tubes, resistors of known temperature, or an oscillator whose frequency is swept through the band at a uniform rate.

c. Comparison Method (*See Section 3.3*): This method consists of direct comparison between the network being tested and a secondary standard in the form of a network of the same type for which the average noise factor has been determined.

Of these three methods the first, involving the direct measurement of noise bandwidth, is complicated. The noise-source dispersed-signal method is simpler, and therefore preferred in both laboratory and production situations. The gas-discharge-generator dispersed-signal method is especially useful for high frequencies where the noise output of a diode may be affected by transit time or lead effects. The swept-oscillator dispersed-signal method may have some application as a supplementary method. Comparison methods are primarily of use in production testing.

The general arrangement of the apparatus for any of these methods is shown in Fig. 1. Certain requirements must be met by each of the three components of Fig. 1.

1) Since the calibrated signal generator is the input termination of the transducer, the behavior of the generator output admittance with frequency, over the pass band of the system, must duplicate that of the input termination with which the measured average noise factor is intended to be associated. The signal generator may deliver either power at a single frequency or power distributed over a frequency spectrum. In the latter case, its available power must be uniform over the pass band of the system. In either case, its available power must be accurately known. The calibration of the generator should take account of any elements that have been added to the basic generator to realize desired output admittance characteristics.

2) The transducer under test must be linear in the sense that its available-output-power change for a given available-input-power change is independent of the initial power level for all values used in testing. It may include linear elements or linear frequency shifters, but, particularly, simple envelope detectors must be excluded unless a noise source with a uniform power spectrum is being used as a signal generator. For example, in testing a conventional heterodyne receiver the power-measuring device must be connected ahead of the second detector unless the latter is itself a suitable power meter meeting the requirements described later.

3) The power-measuring device must indicate quantitatively the relative values of a) the noise power at the output of the system with no signal input, and b) the total power at the output when an input signal is applied. The measuring device may be either a true power-measuring device, such as a bolometer or thermocouple, or some other type of instrument that has been calibrated to read power for the particular wave forms used in testing.

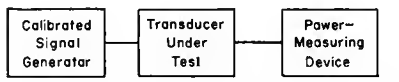

Fig. 1—Average-noise-factor test arrangement.

3.1 *CW-Signal-Generator Method*

The measurement of *average noise factor* with a CW sine-wave signal generator will now be described. The *average noise factor* to be determined may be written

$$\overline{F} = \frac{N_0}{kT_0 \int G(f)df}, \tag{3}$$

where

$G(f)$ = transducer gain at frequency f,
N_0 = noise-power output in watts,
k = Boltzmann's constant, 1.38×10^{-23} joules per degree K,
T_0 = standard noise temperature, 290°K.

[4] The Boulder Laboratories of the National Bureau of Standards have initiated a calibration service for X-band microwave noise sources, and intend to extend this to other common microwave bands as rapidly as possible. Calibration is in terms of the noise power delivered by the terminals of the customer's source to a matched waveguide of standard dimensions for the appropriate microwave frequency band. This power is expressed in db above kT_0B, with an accuracy of 0.1 db. Service is available at three frequencies within the band.

It is convenient to introduce quantities G_0 and B such that

$$G_0B \equiv \int G(f)df, \tag{4}$$

where

G_0 = transducer gain at some convenient reference frequency f_0,

B = noise bandwidth of the system in cycles per second (cps),

$G(f)$ = transducer gain at frequency f.

In general, B is a function of f_0.

With the signal generator connected to the transducer, but with the generator output at zero except for its thermal noise at standard temperature, let the transducer-power output be P_1. With the available signal power of the generator set at P_s at the reference frequency f_0, let the transducer-power output be P_2. Then the transducer gain at f_0 is $G_0=(P_2-P_1)/P_s$. Also the noise-power output will correspond to the original value P_1, so that (3) becomes

$$\overline{F} = \frac{1}{\dfrac{P_2}{P_1}-1}\,\frac{P_s}{kT_0B}\,. \tag{5}$$

To measure $\overline{F}$ accurately it is desirable to choose P_s so that P_2 is several times greater than P_1. However, for convenience and to avoid saturation, it is common to choose P_s so that $P_2=2P_1$. Eq. (5) then becomes $\overline{F}=P_s/kT_0B$, and P_s may be considered to be a power equivalent to the noise output as referred to the input circuit. It will be noted that, since only the ratio P_2/P_1 enters $\overline{F}$, an absolute calibration of the power-output measuring meter is not necessary, although power ratios must be determined accurately. Alternatively, a calibrated attenuator may be employed between the output of the transducer under test and the power-output indicator.

3.1.1 Determination of Noise Bandwidth: The noise bandwidth may be determined from a plot, on linear scales, of the curve of relative transducer gain vs frequency, as shown in Fig. 2. The noise bandwidth is the width, along the frequency axis, of a rectangular response curve with an area M and a height h at f_0 which are the same as those of the actual response curve. The frequency f_0 is usually, but not always, the frequency of maximum response. The area M and height h are those above the noise contribution to the power output, since only the signal output is of interest here. If the area is M, if the height at f_0 is h, and if the frequency scale is D cps per unit of length, the noise bandwidth B referred to f_0 is MD/h. Since the average noise factor refers only to the principal response of the system, spurious or image responses should be excluded from the response curve.

If the characteristics of the network are such that signals large compared with the noise can be handled without saturation, the effect of noise can be ignored in determining B. That is, the P_1 level in Fig. 2 may become negligibly low. It may be possible to achieve this same effect by reducing the gain of the network to considerably less than its normal value, but it should be ascertained that this does not cause undesirable saturation effects, and that bandwidth is not altered by reducing the gain, as may happen if regenerative effects exist.

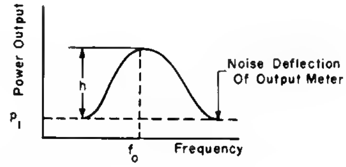

Fig. 2—Transducer relative power output as a function of the frequency of a sine-wave input from a signal generator with fixed available power output.

3.2 Dispersed-Signal-Source Method

Determinations of *average noise* factor may conveniently be made with a signal generator whose available power is distributed uniformly over the response band of the system. This method may eliminate the need for a direct determination of the noise bandwidth, as will be shown below. It is necessary that the available power of the generator in watts per cps of bandwidth be accurately calibrated. If the system is connected as in Fig. 1 with the generator output zero except for the thermal noise at standard temperature, a reading P_1 will be obtained on the power-output meter. (If desired, an equivalent passive termination may be substituted for the generator to obtain P_1.) If the generator is now made to have an available power density of $\mathcal{P}$ watts per cps in addition to the initial thermal noise, the power output will be increased by $\mathcal{P}BG_0$ and the new output-meter reading will be P_2 or, in terms of the same symbols as used previously, $G_0=(P_2-P_1)/\mathcal{P}B$. Also, as before, $N_0=P_1$, or by substitution in (3),

$$\overline{F} = \frac{1}{\dfrac{P_2}{P_1}-1}\,\frac{\mathcal{P}}{kT_0}\,. \tag{6}$$

If the temperature T of the internal impedance of the generator (with the generator output zero except for the thermal noise) is not equal to the standard temperature T_0, (6) does not give the correct noise-figure expression and a correction has to be introduced:

$$\overline{F} = \left(1-\frac{T}{T_0}\right)+\frac{1}{\dfrac{P_2}{P_1}-1}\,\frac{\mathcal{P}}{kT_0}\,. \tag{6a}$$

It should be noted that this measurement may involve spurious or image responses, which will ordinarily give an average noise factor that is deceptively good, *i.e.*, too low, unless appropriate correction is made. In

other words, if the dispersed signal covers pass-band responses not ordinarily used, such as images, which have appreciable gains compared to that of the pass band ordinarily used, the test signal will produce an effect in the output circuit greater than would be produced if the dispersed-signal source were limited so as to include only the response ordinarily used. Hence, if important spurious responses exist, a bandwidth measurement must be made and the *average noise factor* initially determined must be increased in the ratio of 1) the noise bandwidth including all spurious responses, to 2) the noise bandwidth when only the desired response is considered.

3.2.1 Noise-Diode Generators: A temperature-limited diode can be used as a noise generator when connected in a circuit such as that shown in Fig. 3(a). This circuit is typical, but many variations are possible. It is assumed here that the frequency-independent resistive load R_s simulates the real part of the desired transducer source admittance Y_s at every frequency in the pass band of the system. The susceptance B_s' plus any stray susceptance associated with the diode and wiring must simulate the susceptive part B_s of Y_s. The equivalent circuit of this array is shown in Fig. 3(b). The *noise-current generator* i_s represents *thermal noise* in Y_s. For each small element of bandwidth Δf, the mean-square value of i_s for R_s at *standard noise temperature* is

$$\overline{i_s^2} = 4kT_0\Delta f/R_s. \tag{7}$$

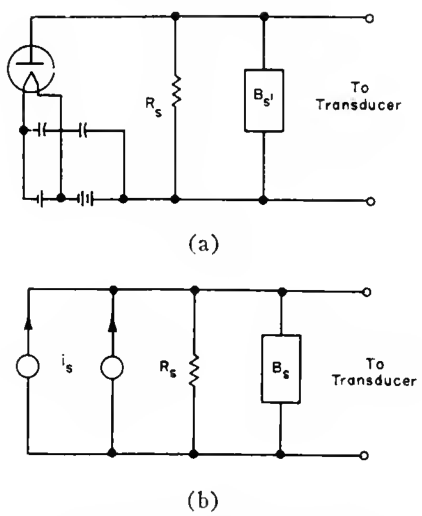

Fig. 3—(a) Typical noise-generator circuit. (b) Equivalent circuit.

The *noise current generator* i_d represents full shot noise in the diode current I_d. If the cathode temperature is adjusted to give a temperature-limited direct plate current of I_d amperes, then, for each small element of bandwidth Δf, the mean-square value of i_d is

$$\overline{i_d^2} = 2eI_d\Delta f, \tag{8}$$

where e is the electronic charge, 1.60×10^{-19} coulomb.

This noise-diode generator circuit has, over and above its thermal noise, a constant available noise power per unit bandwidth of $\mathcal{P}$ watts per cps, where

$$\mathcal{P} = \frac{eI_dR_s}{2}. \tag{9}$$

Inserting (9) into (6), one obtains for the average noise factor the expression

$$\overline{F} = \frac{1}{\left(\dfrac{P_2}{P_1} - 1\right)} \frac{eI_dR_s}{2kT_0}. \tag{10}$$

Since the numerical value of $kT_0/e = 0.0250$ volt,

$$\overline{F} = 20I_dR_s \frac{1}{\left(\dfrac{P_2}{P_1} - 1\right)}, \tag{11}$$

where I_d is in amperes, R_s is in ohms, and P_2/P_1 is the ratio of noise-power outputs.

As before, it is desirable that P_2 be considerably larger than P_1 (preferably several times larger) so that the difference between P_2/P_1 and unity can be determined accurately. For convenience, P_2 is often made equal to $2P_1$. Sometimes a smaller value of P_2 must be used because of limitations imposed on the diode plate current I_d. In such cases additional care must be taken to insure that P_1 and P_2 are stable, repeatable readings to assure reasonable accuracy in the result.

In the foregoing discussions no account has been taken of electron transit time in the diode. If the transit time is an appreciable fraction of a cycle, the noise output of the diode will be lowered.[5] At low frequencies the noise output of diodes may increase above the value (8) because of flicker noise.

3.3 Comparison Methods of Noise Measurement

The methods just described may not be convenient or necessary for production testing. In such cases, noise factors can be checked approximately by carefully chosen comparisons with the performance of a master standard unit of known noise factor. A measurement of SNR after detection, with a fixed modulated-signal input, may be used for checking individual units, provided that the bandwidths of the networks, both preceding and following the detector, and also the detector characteristics are maintained within close limits.

3.4 Precautions

Care should be taken to insure that the appratus attached to the network to be measured does not materially affect the bandwidth of the system. For example, if regeneration is introduced that greatly reduces the bandwidth, the noise factor may be markedly

[5] For transit-time correction, see D. B. Fraser, "Noise spectrum of temperature-limited diodes," *Wireless Engr.*, vol. 26, pp. 129–131; April, 1949.

altered. In any event, undesired feedback usually makes the measurement much more difficult.

Careful shielding and filtering of input and output elements is essential, particularly when the difference in power level between output and input is large. If the measuring-signal generator has a temperature different from standard temperature, an appropriate correction must be applied.

4. Measurement of Spot-Noise Parameters

The noise factor discussed in Section 3 is the weighted average noise factor over the entire pass band of the network. The *spot noise factor*, which is the noise factor at a particular frequency, can be determined by including a very-narrow-band filter of the desired frequency between the network and the power-measuring device. When the spot frequency is near the center of the band, the factor so obtained may not be greatly different from the average.

4.1 Noise Factor of Transducers in Cascade

Frequently, several networks are connected in cascade, and it is desirable to know how the noise factor of each affects the noise factor of the over-all system. This is necessary both in evaluating the effect of improvement in any part of the system and in measuring the noise factor of a single unit in a system.

For a number of networks in cascade, as shown in Fig. 4, the system spot noise factor is given in terms of the component spot noise factors by

$$F = F_1 + (F_2 - 1)/G_1 + (F_3 - 1)/G_1G_2 + \cdots, \quad (12)$$

where G_1, G_2, and so forth, are the available power gains of the component networks.[6]

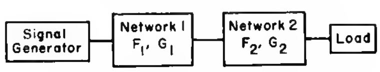

Fig. 4—Networks in cascade.

The spot noise factor F_i and the available gain G_i of the ith network are those obtained with a source impedance equal to the impedance presented by the output of the preceding part of the cascade. An analogous expression for average noise factor can be derived, but is more involved. When the frequency for spot noise factor is chosen near the center of the noise-transmission characteristic, (12) for spot noise factor is often a satisfactory approximation for average noise factor.

These formulas also apply to networks that attenuate rather than amplify, in which case the corresponding gains are less than unity. It should be mentioned that the spot noise factor of a passive attenuating network at standard temperature is the factor by which the available power is attenuated in passing through it. It should also be noted that, when each of the various networks has substantial gain and the noise factor of the later stages is not excessive, then the over-all noise factor is largely determined by the noise factor of the first stage as is obvious from (12).

4.2 The Noise Parameters F_0, G_0, B_0, and R_n

As stated in Section 2, the *noise factor* F of any linear transducer at a given input frequency varies with the admittance $Y_s = G_s + jB_s$ of its input termination in the manner shown in (1), which can be expanded as follows:

$$F = F_0 + \frac{R_n}{G_s}\left[(G_s - G_0)^2 + (B_s - B_0)^2\right], \quad (13)$$

where G_0 and B_0 are the conductive and susceptive parts, respectively, of the optimum source admittance Y_0 cited in Section 2.

The four parameters F_0, G_0, B_0, and R_n characterize the noise properties of the transducer and are independent of the source admittance Y_s. These noise parameters can be determined empirically by fitting this four-parameter expression to observed values of F as a function of Y.

A suitable program for determining F_0, G_0, B_0, and R_n from noise-factor measurements is the following:

1) With the source conductance G_s maintained constant, measure several values of the *noise factor* F for different values of the source susceptance B_s. Plot the curve F vs B_s and determine the optimum source susceptance B_0.
2) With the source susceptance B_s maintained at its optimum value B_0, measure several values of the *noise factor* F for different values of the source conductance G_s. Plot the curve F vs G_s, and determine the optimum source conductance G_0.
3) Using the data already obtained, plot F vs x, where x is the quantity $|Y_s - Y_0|^2/G_s$. These data should lie on a straight line of the form $F = F_0 + R_n x$. The slope of this line is the resistance parameter R_n, and its F intercept at $x = 0$ is the optimum noise factor F_0.

If a direct-reading noise-factor meter is available, the noise-factor minima in Steps 1) and 2) can be observed directly on this meter, yielding values for B_0 and G_0. Additional data of F vs x for Step 3) can then be obtained by using other convenient values of Y_s.

If the points in Step 3) do not line on a straight line, then

a) The estimates of B_0 and G_0 in Steps 1) and 2) may be inaccurate,[7]
b) The individual noise-factor measurements may be in error, or
c) The transducer under test may be nonlinear.

[6] If a negative output resistance exists somewhere in the cascade, the noise factor of the over-all system can be calculated as described by H. A. Haus and R. B. Adler, "An extension of the noise figure definition," Proc. IRE, vol. 45, pp. 690–691; May, 1957.

[7] This can occur if the minima in the F-vs-B_s and -G_s curves are very shallow. In this case, slightly adjusted values of B_0 and G_0 may improve the linearity of the F-vs-x curve.

4.2.1 Equipment: The measuring equipment for determining the noise factor is discussed in Section 3. For the measurement of the noise parameters, it is necessary to provide, in addition, a means for adjusting the source admittance. The type of equipment needed depends on the frequency of the measurement.

An arrangement suitable for use at frequencies below 100 or 200 Mc is shown in Fig. 5. The susceptance is adjustable by means of a calibrated capacitance or inductance, and the conductance is varied by using different resistances or a variable resistance. At the higher frequencies of this range, care must be taken that the susceptance B_s is maintained at its optimum value B_0 when the conductance G_s is varied in the measurements specified in Step 2). A temperature-limited diode or a resistor at known temperature can be used as a noise source. Gas discharge tubes combined with calibrated attenuators have also been used in the upper part of this frequency range.

Noise generators suitable for use at higher frequencies are, by nature of their construction, devices with fixed output conductance. A calibrated source with variable internal admittance can be made by combining such a generator with a network that acts like a variable-ratio transformer. An arrangement used for measurements on amplifiers at frequencies above 500 Mc is shown in Fig. 6.

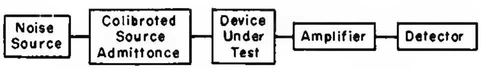

Fig. 5—Test arrangement for determination of noise parameters.

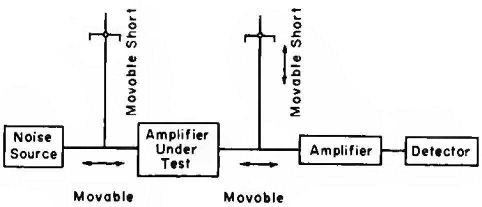

Fig. 6—Test arrangement used for measurements on amplifiers at frequencies above 500 Mc.

The transformer consists of a section of transmission line fitted with an adjustable stub that can be moved along the length of the line. Two identical units are used. One unit couples the amplifier to the noise generator or signal generator. The second unit couples the output of the amplifier to the succeeding stage and is used to facilitate the measurement by maximizing the output. The correction due to the noise of the second stage is correspondingly reduced.

4.2.2 Sample Determination of the Noise Parameters at Low Frequencies: The determination of the noise parameters of a germanium alloy junction *p-n-p* transistor, operated in the common-emitter connection at a frequency of 900 cps, is presented in this section.

Fig. 7 shows observed F-vs-B_s data for a constant source conductance of 1.00 millimho. The optimum source susceptance B_0 is zero. Fig. 8 shows observed F-vs-G_s data for $B_s = B_0 = 0$. The optimum source conductance G_0 is 0.5 millimho. Fig. 9 shows the F-vs-x plot obtained from the data used previously in Figs. 7 and 8. The intercept and slope of the line plotted by use of these data yield $F_0 = 1.55$ and $R_n = 540$ ohms. The curves superposed on the experimental data in Figs. 7 and 8 were computed from (13) for the values of the noise parameters F_0, G_0, B_0, and R_n determined above.

4.2.3 Sample Determination of the Noise Parameters at High Frequencies: The determination of the noise parameter for a disk-seal triode, made at a frequency of 870 Mc, is described in this section. The arrangement

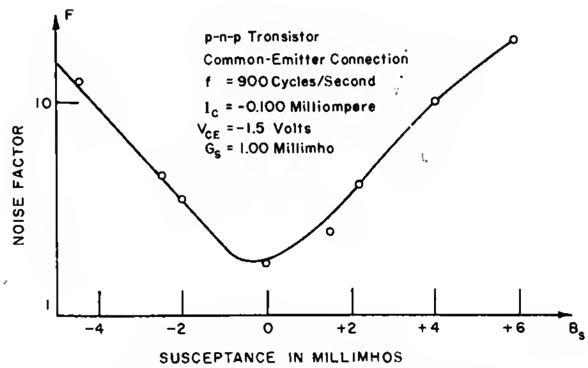

Fig. 7—Noise factor vs source susceptance.

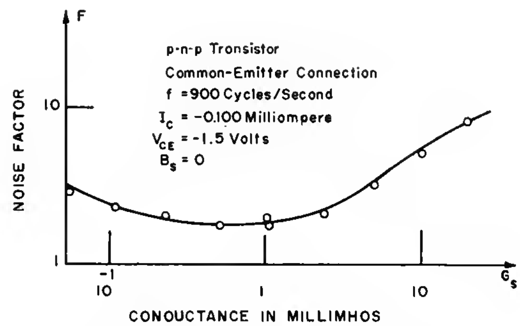

Fig. 8—Noise factor vs source conductance.

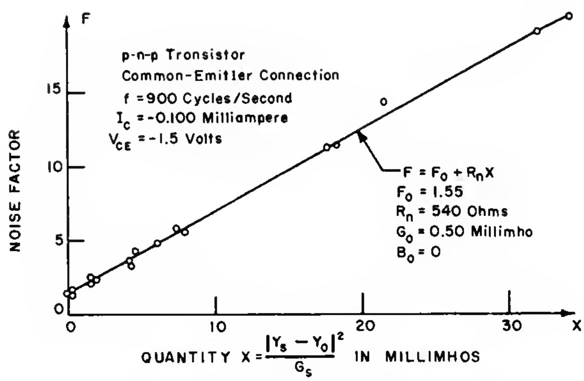

Fig. 9—Noise factor vs quantity x.

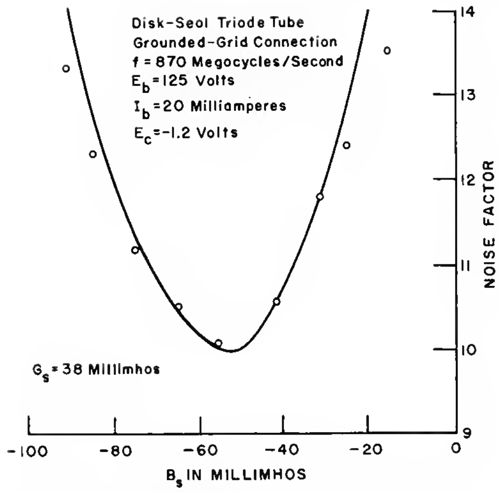

Fig. 10—Noise factor vs source susceptance.

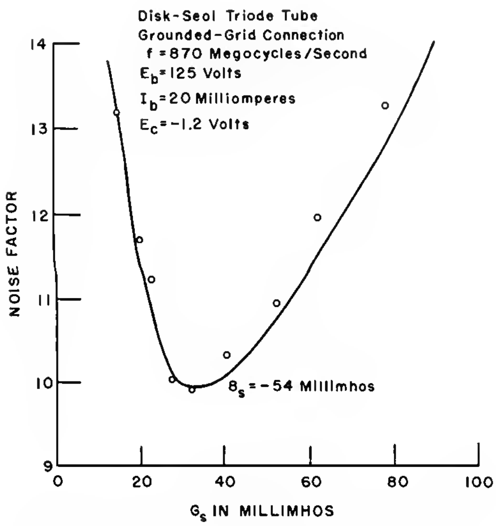

Fig. 11—Noise factor vs source conductance.

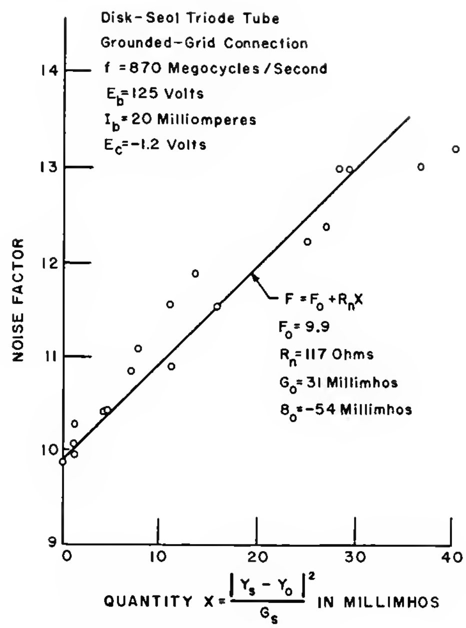

Fig. 12—Noise factor vs quantity x.

of equipment shown in Fig. 6 was used. A series of readings with input-stub position and length chosen to give $G_s = 38$ millimhos for various values of B_0 provided the data shown in Fig. 10. The optimum source susceptance B_0 is found to be -54.0 millimhos. Fig. 11 shows observed F-vs-G_s data for $B_s = B_0 = -54.0$ millimhos. The optimum source conductance is 32.0 millimhos. Fig. 12 shows the F-vs-x plot obtained from the data used in Figs. 10 and 11. The intercept and slope of the line drawn by use of these data yield $F_0 = 9.9$ and $R_n = 117$ ohms. The curves superposed on the experimental data in Figs. 10 and 11 were computed from (13) by using the values of the noise parameters F_0, G_0, B_0 and R_n determined above.

APPENDIX

Representation of Noise in Linear Twoports

Summary—**This is a tutorial paper, written by the Subcommittee on Noise, IRE 7.9, to provide the theoretical background for the preceding sections of Part 9—Noise in Linear Twoports. The general-circuit-parameter representation of a linear twoport with internal sources and the Fourier representations of stationary noise sources are reviewed. The relationship between spectral densities and mean-square fluctuations is given, and the noise factor of the linear twoport is expressed in terms of the mean-square fluctuations of the source current and the internal noise sources. The noise current is then split into two components, one perfectly correlated and one uncorrelated with the noise voltage. Expressed in terms of the noise voltage and these components of the noise current, the noise factor is then shown to be a function of four parameters which are independent of the circuit external to the twoport.**

1. Introduction

One of the basic problems in communication engineering is the distortion of weak signals by the ever-present thermal noise and by the noise of the devices used to process such signals. The transducers performing signal processing such as amplification, frequency mixing, frequency shifting, etc., may usually be classified as twoports. Since weak signals have amplitudes small compared with, say, the grid-bias voltage of a vacuum tube, or emitter-bias current of a transistor etc., the amplitudes of the excitations at the ports can be linearly related. Consequently, the description and measurement of noise that is presented in this paper, and which forms the basis for the methods of measurement described in the preceding sections, can be restricted to linear noisy twoports. Even if inherently nonlinear characteristics are involved, as in mixing, etc., linear relations still exist among the signal voltage and input and output amplitudes, although these may not be associated with the same frequency. Image-frequency components may be eliminated by proper filtering, but a generalization of our results to cases in which image frequencies are present is not difficult.

The effect of the noise originating in a twoport when the signal passes through the twoport is characterized at any particular frequency by the *(spot) noise factor (noise figure)*. Since this has meaning only if the source impedance used in obtaining the noise factor is also specified, the noise contribution of a twoport is often indicated by a minimum noise factor and the source impedance with which this minimum noise factor is achieved. However, the extent to which the noise factor depends upon the input source impedance, upon the amount of feedback, etc., can be indicated only by a more detailed representation of the linear twoport.[1,2]

[1] H. Rothe and W. Dahlke, "Theory of noisy fourpoles," Proc. IRE, vol. 44, pp. 811–818; June, 1956. Also, "Theorie rauschender Vierpole," *Arch. elekt. Übertragung*, vol. 9, pp. 117–121; March, 1955.

[2] A. G. T. Becking, H. Groendijk, and K. S. Knol, "The noise factor of four-terminal networks," *Philips Res. Repts.*, vol. 10, pp. 349–357; October, 1955.

As a basis for understanding the representation of networks containing (statistical) noise sources, we shall consider first the analysis of networks containing Fourier-transformable signal sources. We shall then describe noise in two ways: a) as a limit of Fourier-integral transforms, and b) as a limit of Fourier-series transforms. The reasons for the wider use of the latter description will be presented. With this background we shall be ready to show that the four noise parameters used in the preceding sections completely characterize a noisy linear twoport.

2. Representations of Linear Twoports

The excitation at either port of a linear twoport can be completely described by a time-dependent voltage $v(t)$ and the time-dependent current $i(t)$. (For wave-guide twoports, where no voltage or current can be identified uniquely, an equivalent voltage and current can always be used.) Let it be assumed that the voltage and current functions can be transformed from the time domain to the frequency domain, and that V and I stand for the Fourier transforms when the function is aperiodic and for the Fourier amplitudes when the function is periodic. The linearity of the twoport *without internal sources* then allows an impedance representation:

$$V_1 = Z_{11}I_1 + Z_{12}I_2$$
$$V_2 = Z_{21}I_1 + Z_{22}I_2. \quad (1)$$

The subscripts 1 and 2 refer to the input and output ports, respectively, and the coefficients Z_{jk} are, in general, functions of frequency. The currents are defined to be positive if the flow is into the network as shown in Fig. 1.

If the twoport contains internal sources, then (1) and the equivalent circuit must be modified. By a generalization of Thévenin's theorem, the twoport may be separated into a source-free network and two voltage generators, one in series with the input port and one in series with the output port. If the time-dependent functions $e_1(t)$ and $e_2(t)$ describing these equivalent generators can be transformed to functions of frequency E_1 and E_2, respectively, the impedance representation of the linear twoport *with internal sources* becomes

$$V_1 = Z_{11}I_1 + Z_{12}I_2 + E_1$$
$$V_2 = Z_{21}I_1 + Z_{22}I_2 + E_2, \quad (2)$$

and the equivalent nework is that shown in Fig. 2.

For the purpose of the analysis to follow, it should be emphasized again that such equations in general characterize the behavior of the twoport as a function of frequency. For practical purposes, it should be pointed

out that in most cases the impedance parameters of a linear twoport can be measured at a particular frequency by applying sinusoidal voltages that produce outputs large compared with those caused by the internal sources.

The impedance representation of a linear twoport with internal sources has been reviewed because of its familiarity. However, it is well known that other representations, each leading to a different separation of the internal sources from the twoport, are possible. A particularly convenient one for the study of noise is the general-circuit-parameter representation[3]

$$V_1 = AV_2 + BI_2 + E$$
$$I_1 = CV_2 + DI_2 + I,$$

where E and I are again functions of frequency which are the Fourier transforms of the time-dependent functions $e(t)$ and $i(t)$ describing the internal sources.

As shown in Fig. 3, the internal sources are now represented by a source of voltage acting in series with the input voltage and a source of current flowing in parallel with the input current. It will be seen that this particular representation of the internal sources leads to four noise parameters that can easily be derived from single-frequency measurements of the twoport noise factor as a function of input mismatch. It has the further advantage that such properties of the twoport as

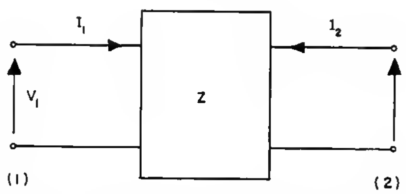

Fig. 1—Sign convention for impedance representation of linear twoport.

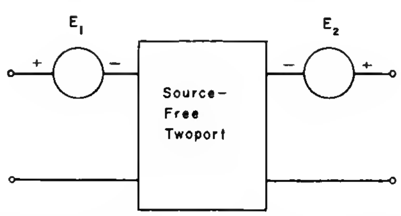

Fig. 2—Separation of twoport with internal sources into a source-free twoport and external voltage generators.

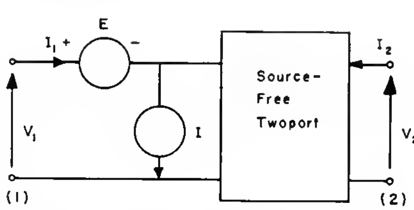

Fig. 3—Separation of twoport with internal sources into a source-free twoport and external input current and voltage sources.

[3] E. Guillemin, "Communication Networks," John Wiley and Sons, Inc., New York, N. Y., vol. 2, p. 138; 1935.

gain and input conductance do not enter into the noise-factor expression in terms of these four parameters.

3. Representations of Stationary Noise Sources

When a linear twoport contains stationary internal noise sources, the frequency functions E and I in (3) cannot be found by conventional Fourier methods from the time functions $e(t)$ and $i(t)$, since these are now random functions extending over all time and have infinite energy content. Two alternatives are possible. One may consider the substitute functions

$$\begin{aligned} e(t, T) &= e(t) \qquad \text{for } -\frac{T}{2} < t < \frac{T}{2} \\ &= 0 \qquad \text{for } \frac{T}{2} < |t| \\ i(t, T) &= i(t) \qquad \text{for } -\frac{T}{2} < t < \frac{T}{2} \\ &= 0 \qquad \text{for } \frac{T}{2} < |t|, \end{aligned} \tag{4}$$

where T is some long but finite time interval, and use the Fourier transforms

$$\begin{aligned} E(\omega, T) &= \frac{1}{2\pi}\int_{-T/2}^{T/2} e(t, T)e^{-j\omega t}dt \\ I(\omega, T) &= \frac{1}{2\pi}\int_{-T/2}^{T/2} i(t, T)e^{-j\omega t}dt. \end{aligned} \tag{5}$$

Alternatively, one may construct periodic functions

$$\begin{aligned} e(t, T) &= e(t) \qquad \text{for } -\frac{T}{2} < t < \frac{T}{2} \\ e(t + nT, T) &= e(t, T) \qquad \text{with } n \text{ an integer,} \\ i(t, T) &= i(t) \qquad \text{for } -\frac{T}{2} < t <, \frac{T}{2} \\ i(t + nT, T) &= i(t, T) \qquad \text{with } n \text{ an integer,} \end{aligned} \tag{6}$$

and expand these functions into Fourier series with amplitudes

$$\begin{aligned} E_m(\omega, T) &= \frac{1}{T}\int_{-T/2}^{T/2} e(t, T)e^{-j\omega t}\, dt \\ I_m(\omega, T) &= \frac{1}{T}\int_{-T/2}^{T/2} i(t, T)e^{-j\omega t}dt, \end{aligned} \tag{7}$$

where $\omega = m2\pi/T$ with m an integer, and T is again finite. In either case, the substitute functions can be made to approach the actual functions as closely as desired by making the interval T larger and larger.

In either the Fourier-integral or the Fourier-series approach, we consider a set of substitute functions obtained in principle from a series of measurements a) on an ensemble of systems with identical statistical properties, or b) on one and the same system at successive

time intervals sufficiently separated so that no statistical correlation exists.[4]

Since noise is a statistical process, we are in general interested in statistical averages[5] rather than the exact details of any particular noise function. These averages are important since they relate to physically measurable stationary quantities. For example, in the Fourier-integral approach the spectral density of the noise-voltage excitation is defined by

$$W_e(\omega) = \lim_{T\to\infty} \frac{\overline{|E(\omega, T)|^2}}{2\pi T}, \tag{8}$$

where the bar indicates an arithmetic average of the Fourier transforms of an ensemble of functions described by (4). This spectral density is proportional to the noise power (in a narrow frequency band[6] around a given frequency) in a resistor across which the fluctuating voltage $e(t)$ appears.

Unfortunately, the notation developed in the literature for dealing with spectral densities is unwieldy, since the quantity with which the density is associated is relegated to a subscript. This is one of the reasons why researchers on noise have tended to use the historically older Fourier-series approach. The use of spectral densities is preferred in rigorous mathematical treatments of noise and in questions involving definitions of noise processes, since this assures that all points on the frequency axis within a narrow range are equivalent. For practical purposes, however, the noise amplitudes that are associated with discrete frequencies in the Fourier-series method can be as closely distributed as desired by choosing the time interval T sufficiently large.

4. Relationship of Spectral Densities and Fourier Amplitudes to Mean-Square Fluctuations

If there is an open-circuit noise voltage $v(t)$ across a terminal pair, the mean-square value $v^2(t)$ is related to the spectral density $W_v(\omega)$ and to the Fourier amplitudes $V_m(\omega)$ as follows:

$$\begin{aligned}\overline{v^2(t)} &= \lim_{T\to\infty}\frac{1}{T}\int_{-T/2}^{T/2} v^2(t)dt \\ &= \int_{-\infty}^{+\infty} W_v(\omega)d\omega = \sum_{m=-\infty} |\overline{V_m(\omega)}|^2, \end{aligned}\tag{9}$$

where the amplitudes V_m are now those obtained as $T\to\infty$.

[4] The equivalence of the ensemble and time average is known as the ergodic hypothesis. An interesting discussion of the implications of this hypothesis can be found in M. Born, "Natural Philosophy of Cause and Chance," Oxford University Press, New York, N. Y., p. 62; 1948. See also, W. B. Davenport, Jr. and W. L. Root, "An Introduction to the Theory of Random Signals and Noise," McGraw-Hill Book Co., Inc., New York, N. Y., p. 66 *ff.*; 1958.

[5] W. R. Bennett, "Methods of solving noise problems," Proc. IRE, vol. 44, pp. 609–637; May, 1956.

[6] A frequency interval Δf is called narrow if the physical quantities under consideration are independent of frequency throughout the interval, and if $\Delta f \ll f_0$; where f_0 is the center frequency.

Suppose that the stationary time function $v(t)$ is passed through an ideal band-pass filter with the narrow bandwidth $\Delta f = \Delta\omega/2\pi$ centered at a frequency $f^0 = \omega^0/2\pi$. The mean-square value of the voltage appearing at the filter output, usually denoted by the symbol $\overline{v^2}$ and called "the mean-square fluctuation of $v(t)$ within the frequency interval Δf" is

$$\overline{v^2} = 4\pi\Delta f\, W_v(\omega_0) = 2\,\overline{|V_m|^2}. \tag{10}$$

This equation relates the mean-square fluctuations, the spectral density, and the Fourier amplitude. If the filter is narrow-band but not ideal, the quantity Δf is the noise bandwidth. The factor 2 in (10) arises because $W_v(\omega)$ and V_m have been defined on the negative as well as the positive frequency axis.

Eq. (10) shows that spectral densities and mean-square fluctuations are equivalent. Although mathematical limits are involved in the definitions, these quantities can be measured to any desired degree of accuracy. For example, if one measures the power flowing into a termination, connected to the terminals with which $v(t)$ is associated, one measures essentially $W_v(\omega_0)$, provided that the following requirements are met:

1) The termination is known and has a high impedance so that the voltage across the terminals remains essentially the open-circuit voltage $v(t)$, or the internal impedance associated with the two terminals is also known so that any change in voltage can be computed. The noise voltage contributed by the termination must either be negligible, or its statistical properties must be known so that its effect can be taken into account.
2) The termination is fed through a filter with a pass band sufficiently narrow so that the spectral density and internal impedance are essentially constant throughout the band; or the termination is itself a resonant circuit with a high Q corresponding to a sufficiently narrow bandwidth.
3) The power measurement is made over a period of time that is long compared to the reciprocal bandwidth $1/\Delta f$ of the filter. In this way the power-measuring device takes an average over many long time intervals that is equivalent to an ensemble average.

In the description of noise transformations by linear twoports that follows, the mean-square fluctuations will be used. Mean-square current fluctuations can be related to the Fourier amplitudes $I_m(\omega)$ by a procedure similar to the one just described. Since fluctuations of the cross products of statistical functions will also be used it should be noted that

$$\lim_{T\to\infty}\frac{1}{T}\int_{-T/2}^{T/2} v(t, T)i(t, T)dt = \sum_{m=-\infty}^{\infty} \overline{V_m I_m^*}, \tag{11}$$

where $i(t)$ is a noise current and V_m and I_m are again the amplitudes obtained as $T\to\infty$.

We may interpret the real part of the complex quantity $2\overline{V_m I_m^*}$ as the contribution to the average of $\overline{vi}$ from the frequency increment $\Delta f = 1/T$ at the angular frequency $\omega = m\Delta\omega$. However, the imaginary part of $V_m I_m^*$ also contains phase information that has to be used in noise computations. We shall, therefore, use the expression

$$\overline{vi^*} = \overline{2V_m I_m^*} \qquad \text{for } m > 0,\ \omega > 0 \tag{12}$$

as the complex cross-product fluctuations of $v(t)$ and $i(t)$ in the frequency increment Δf. They are related to the cross-spectral density by

$$\overline{vi^*} = 4\pi\Delta f\, W_{iv}(\omega), \tag{13}$$

where

$$W_{iv}(\omega) = \lim_{T\to\infty} \frac{\overline{I^*(\omega, T)V(\omega, T)}}{2\pi T}. \tag{14}$$

There are several ways of specifying the fluctuations (or the spectral densities) that characterize the internal noise sources. A mean-square voltage fluctuation can be given directly in units of volt² second. It is often convenient, however, to express this quantity in resistance units by using the Nyquist formula, which gives the mean-square fluctuation of the open-circuit noise voltage of a resistor R at temperature T as

$$\overline{e^2} = 4kTR\Delta f,$$

where k is Boltzmann's constant. For any mean-square voltage fluctuation $\overline{e^2}$ within the frequency interval Δf, one defines the equivalent noise resistance R_n as

$$R_n = \frac{\overline{e^2}}{4kT_0\Delta f}, \tag{15}$$

where T_0 is the standard temperature, 290°K. The use of R_n has the advantage that a direct comparison can be made between the noise due to internal sources and the noise of resistances generally present in the circuit. Note that R_n is not the resistance of a physical resistor in the network in which e is a physical noise voltage and therefore does not appear as a resistance in the equivalent circuit of the network.

In a similar manner, a mean-square current fluctuation can be represented in terms of an equivalent noise conductance G_n which is defined by

$$G_n = \frac{\overline{i^2}}{4kT_0\Delta f}. \tag{16}$$

5. Noise Transformations by Linear Twoports

The statistical averages discussed in Sections 3 and 4 will now be used to describe the internal noise sources of a linear twoport. We may consider functions of the type given by (6) and represent the noise sources E and I in (3) by the Fourier amplitudes $E_m(\omega, T)$ and $I_m(\omega, T)$. Thus, if the circuit shown in Fig. 3 represents a separation of noise sources from a linear twoport, the noise-free circuit is preceded by a noise network. Since a noise-free network connected to a terminal pair does not change the SNR (noise factor evaluated at that terminal pair) the noise factor of the over-all network is equal to that of the noise network.

To derive the noise factor, let us connect the noise network to a statistical source comprising an internal admittance Y_s and a current source again represented by a Fourier amplitude $I_s(\omega, T)$.[7] The network to be used for the noise-factor computation is then as shown in Fig. 4. By definition, the *spot noise factor* (*figure*) of a network at a specified frequency is given by the ratio of 1) the total output noise power per unit bandwidth available at the output port to 2) that portion of 1) engendered by the input termination at the standard temperature T_0.

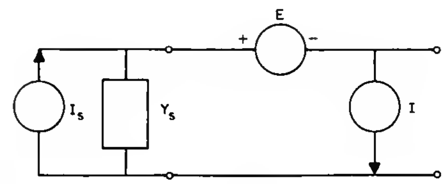

Fig. 4—Truncated network for noise-factor computation.

Now the total short-circuit noise current at the output of the network shown in Fig. 4 can be represented in terms of Fourier amplitudes by

$$-I_{sm}(\omega, T) + I_m(\omega, T) + Y_s E_m(\omega, T).$$

Let us assume that the internal noise of the twoport and the noise from the source are uncorrelated. If we then square the total short-circuit noise-current Fourier amplitudes, take ensemble averages and use equations similar to (10) and (12) to introduce mean-square fluctuations, we obtain a mean-square current fluctuation

$$\overline{i_s^2} + \overline{|i + Y_s e|^2} = \overline{i_s^2} + \overline{i^2} + |Y_s|^2\overline{e^2} + Y_s^*\overline{ie^*} + Y_s\overline{i^*e} \tag{17}$$

to which the total output noise power is proportional. It should be noted here that, when e and i are complex, the symbols $\overline{e^2}$ and $\overline{i^2}$ denote, by convention, $\overline{|e|^2}$ and $\overline{|i|^2}$. Since the noise power due to the source alone is proportional to $\overline{i_s^2}$, the noise factor becomes

$$F = 1 + \frac{\overline{|i + Y_s e|^2}}{\overline{i_s^2}}. \tag{18}$$

In the denominator, the mean-square source noise current is related to the source conductance G_s by the Nyquist formula

$$\overline{i_s^2} = 4kT_0 G_s \Delta f. \tag{19}$$

[7] "IRE Standards on Electron Tubes: Definitions of Terms, 1957," (57 IRE S.2) Proc. IRE, vol. 45, pp. 983–1010; July, 1957.

In the numerator of (18) the four real variables involved in $\overline{i^2}$, $\overline{e^2}$ and $\overline{ie^*}$, where e^* is the complex conjugate of e, describe the internal noise sources. If the Fourier transforms (5) are used to characterize the noise, the self-spectral densities of noise current and voltage and the cross-spectral density of these quantities would have to be specified.

Before proceeding, let us review in general terms the methods employed and the conclusions reached. The representation of the internal noise sources of a noisy linear twoport by a voltage generator and a current generator lumps the effect of all internal noise sources into the two generators. A complete specification of these generators is thus equivalent to a complete description of the internal sources, as far as their contribution to terminal voltages and currents is concerned. Since the sources under consideration are noise sources, their desciption is confined to the methods applicable to noise. The extent and detail of the description depends on the amount of detail envisioned in the analysis. Since, in the case of the noise factor, only the mean-square fluctuations of output currents are sought, the specification of self and cross-product fluctuations of the generator voltages and currents is adequate.

The expression for the noise factor given in (18) can be simplified if the noise current is split into two components, one perfectly correlated and one uncorrelated with the noise voltage. The uncorrelated noise current, designated by i_u, is defined at each frequency by the relations

$$\overline{ei_u^*} = 0, \tag{20}$$

$$\overline{(i - i_u)i_u^*} = 0. \tag{21}$$

The correlated noise current $i - i_u$ can be written as $Y_\gamma e$, where the complex constant $Y_\gamma = G_\gamma + jB_\gamma$ has the dimensions of an admittance and is called the correlation admittance. The cross-product fluctuation $\overline{ei^*}$ may then be written

$$\overline{ei^*} = \overline{e(i - i_u)^*} = Y_\gamma^* \overline{e^2}. \tag{22}$$

The noise-voltage fluctuation can be expressed in terms of an equivalent noise resistance R_n as

$$\overline{e^2} = 4kT_0R_n\Delta f, \tag{23}$$

and the uncorrelated noise-current fluctuation in terms of an equivalent noise conductance G_u as

$$\overline{i_u^2} = 4kT_0G_u\Delta f. \tag{24}$$

The fluctuations of the total noise current are then

$$\begin{aligned}\overline{i^2} &= \overline{|\, i - i_u \,|^2} + \overline{i_u^2} \\ &= 4kT_0[\,|\, Y_\gamma \,|^2 R_n + G_u]\Delta f.\end{aligned} \tag{25}$$

From (18)–(24), the formula for the noise factor becomes

$$\begin{aligned}F &= 1 + \frac{1}{4kT_0G_s\Delta f}\left[\overline{i_u^2} + |\, Y_s + Y_\gamma \,|^2 \overline{e^2}\right] & (26)\\ &= 1 + \frac{G_u}{G_s} + \frac{R_n}{G_s}[(G_s + G_\gamma)^2 + (B_s + B_\gamma)^2]. & (27)\end{aligned}$$

Thus, the noise factor is a function of the four parameters G_u, R_n, G_γ and B_γ. These depend, in general, upon the operating point and operating frequency of the twoport, but not upon the external circuitry. In a vacuum-tube triode the correlation susceptance B_γ is negligibly small at frequencies such that transit times are small compared to the period. Also, G_u and G_γ are vanishingly small when there is little grid loading. Thus, tube noise at low frequencies is adequately characterized by the single nonzero constant R_n. Tube noise at high frequencies and transistor noise at all frequencies have no such simple representation.

Since the noise factor is an explicit function of the source conductance and susceptance, it can readily be shown that the noise factor has an optimum (minimum) value at some optimum source admittance $Y_0 = G_0 + jB_0$, where

$$G_0 = \left[\frac{G_u + R_nG_\gamma^2}{R_n}\right]^{1/2} \tag{28}$$

$$B_0 = -B_\gamma, \tag{29}$$

and the value of this minimum noise factor is

$$F_0 = 1 + 2R_n(G_\gamma + G_0). \tag{30}$$

In terms of G_0, B_0 and F_0, the noise factor for any arbitrary source impedance then becomes

$$F = F_0 + \frac{R_n}{G_s}[(G_s - G_0)^2 + (B_s - B_0)^2]. \tag{31}$$

Eq. (31) shows that the four real parameters F_0, G_0, B_0 and R_n give the noise factor of a twoport for every input termination of the twoport.[8] As shown in the methods of measurement described in the preceding sections, a measurement of the minimum noise factor and of the source admittance Y_0 with which F_0 is achieved gives the first three parameters. The parameter R_n can be computed from an additional measurement of the noise factor for a source admittance Y_s other than Y_0.

From the given values F_0, G_0, B_0 and R_n, one can compute, if desired, the noise fluctuations $\overline{i^2}$, $\overline{e^2}$, and $\overline{ei^*}$ (or the corresponding spectral densities). To do this, one uses (28)–(30) to find Y_γ and G_u. From these one evaluates $\overline{e^2}$, $\overline{i^2}$, and $\overline{ei^*}$ from (23), (25), and (22). The

[8] Reasoning similar to that leading to (31) can be carried out in a dual representation, where impedances and admittances are interchanged. An equation similar to (31) results, again involving four noise parameters, some of which are different from the ones used here.

fluctuations of any terminal voltage or current of the twoport produced with a given source and load can then be evaluated from the known coefficients (A, B, C, D) of (3), and the known noise fluctuations. Thus, the noise in a twoport is completely characterized (with regard to the fluctuations or the spectral densities at the terminals) by the noise fluctuations $\overline{e^2}$, and $\overline{i^2}$, $\overline{ei^*}$, or alternately by the four noise parameters F_0, G_0, B_0 and R_n.

6. Conclusion

The preceding discussion showed that, with limited objectives, the noise in a linear twoport can be characterized adequately by a limited number of parameters. Thus, if one seeks only information concerning the mean-square fluctuations or the spectral densities of currents or voltages into or across the ports of a linear twoport at a particular frequency and for arbitrary circuit connections of the twoport, it is sufficient to specify two mean-square fluctuations and one product fluctuation or two spectral densities and one cross-spectral density. This involves the specification of four real numbers. If a band of frequencies is being considered, these quantities have to be given as functions of frequency unless they are approximately constant over the band.

Different separations of internal noise sources will lead, in general, to different frequency dependences of the resulting fluctuations or spectral densities. Thus, a particular separation of the noise sources may be preferable to another separation if it is found that fluctuations (spectral densities) are less frequency-dependent in the band. In particular, if available information about the physics of the noise in a particular device suggests introducing, *inside the twoport*, appropriate noise generators of fluctuations (spectral densities) with no frequency dependence, or with a simple frequency dependence, it may be more advantageous to specify the device in terms of these physically suggestive generators.

However, usually one resorts to the characterizations of noise presented in this paper, since they have the advantage that they do not require any knowledge of the physics of the internal noise. Furthermore, noise-factor measurements performed on the twoport yield (more or less) directly the noise parameters of the general circuit representation, namely the fluctuations (spectral densities) of the voltage and current generators attached to the input of the noise-free equivalent of the twoport. This fact recommends the use of the noise parameters natural to the general-circuit-parameter representation.

Part 10

Cathode-Ray Charge Storage Tubes

Subcommittee 7.10

Storage Tubes

A. S. Luftman, *Chairman* (1955–1962)

H. Albertine	D. Davis	J. A. McCarthy
C. J. Andrasco	B. E. Day	W. R. Miller
J. A. Buckbee	H. J. Evans	W. E. Mutter
A. E. Beckers	M. D. Harsh	D. S. Peck
A. Bramley	H. O. Hook	D. L. Schaefer
J. Burns	R. B. Janes	R. W. Sears
G. Chafaris	A. S. Jensen	H. M. Smith
H. G. Cooper	N. J. Koda	M. M. de Thomasson
C. L. Corderman	B. Kazan	W. O. Unruh
M. Crost	M. Knoll	S. T. Yanagisawa
F. R. Darne	C. C. Larson	P. Youtz

1. Introduction

A cathode-ray charge storage tube is a charge storage tube wherein the information is introduced by means of a cathode-ray beam. The information may be read out at a later time, with the output either an electrical signal and/or a visual image corresponding to the stored charge pattern. Storage-tube types using other input techniques than a cathode-ray beam, for example, camera tubes, will not be considered.

Many tests specified for cathode-ray tubes are applicable to cathode-ray charge storage tubes and should be considered in addition to the tests outlined in this part.

Several tests to be described require that there be stored on the target a *standard raster*, defined as a rectangular pattern composed of uniformly-spaced lines scanned at a constant speed.

Another group of tests to be described requires that a *standard array* of spots be used which is square in outline and detail. This array is generated by using staircase deflection waveforms. (The results obtained using random-access deflection may not be identical with those obtained using staircase deflection.)

Operating conditions not specified for a particular test should remain constant throughout the test series.

For purposes of this standard, beam current means the current emerging from the final apertures of the electron gun.

As in any test, care must be taken to insure that the measuring equipment is adequate for the test. In particular, the bandwidth of amplifiers in the test equipment must be adequate for accurate resolution measurements.

2. Types of Cathode-Ray Charge Storage Tubes

For test purposes storage tubes are frequently classified according to a) the types of input to and output derived from the tubes, and b) whether the beam is scanned or indexed during writing. The following classes are used throughout this part.

2.1 Classifications by Output

Cathode-ray charge storage tubes are classed as electrical-signal storage tubes where the output of the tube is an electrical signal or as electrical-visual storage tubes where the output of the tube is a visual display.

2.2 Classifications by Deflection Pattern

The tubes are classed as beam-indexing storage tubes where the writing electron beam is indexed to discrete positions on the storage target or as scanned storage tubes where writing is accomplished during the continuous scanning of the storage surface with an electron beam.

2.3 Combination Storage Tubes

Many storage tube types can be operated with either beam-indexing or scanned writing. Such tubes may be tested under each classification or under the classification used in the case of a specific application.

3. Measurement of Resolution

The quantity of information which can be stored and distinguished in the output, related to resolution, may be limited in a charge storage tube by such factors as the

beam spot size, physical characteristics of the storage area and charge redistribution on the storage surface.

The recommended test techniques for determining resolution quantitatively are different for the several classes of storage tubes described below.

3.1 Resolution of Scanned Electrical-Signal Storage Tubes

Under stated operating conditions a standard raster is written once and read orthogonally. The peak-to-peak amplitude of the read signal is independent of line spacing at large spacings but decreases as the spacing decreases. The amplitude of peak-to-peak output signal, relative to the peak-to-peak output signal for large line spacings, is determined as a function of the number of lines written across the diameter of the storage surface (or the diagonal of the storage surface in the case of tubes having rectangular surfaces). A complete specification of resolution is a curve of number of lines written per diameter or diagonal vs relative output amplitude. (See Fig. 1.) The number of lines writter per diameter or diagonal corresponding to a relative amplitude of 0.5 (called half-amplitude resolution) may be used as a single value for resolution. When "TV lines" resolution is specified, both lines and spaces are counted, giving a value twice that obtained above.

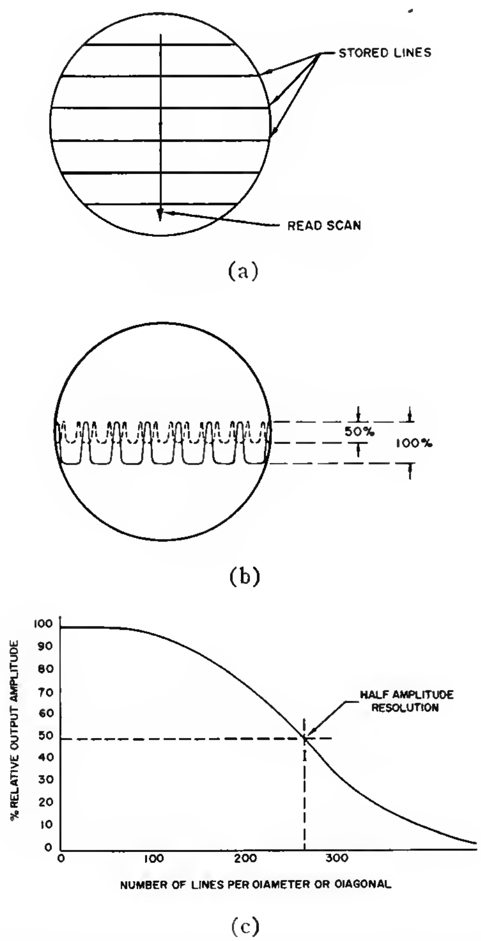

Fig. 1—Resolution measurements. (a) Storage surfaces. (b) Oscilloscope picture of relative output amplitude. (c) Typical resolution plot.

The resolution specification should include separate measurements taken in orthogonal directions with identical tube operating conditions. To permit analysis of deflection defocusing and aberrations, a complete tube specification should also specify the half-amplitude resolution under the following conditions:

a) The resolution at the center of the storage area with the beam focused at the center.
b) The resolution along a diameter or diagonal, at a point 90 per cent of the distance from the center to the edge with the beam focused at the center.
c) The resolution at the point specified in b) with the beam focused at that point.

In making the above measurements, care should be taken to insure that the writing speed remains constant throughout, that the beam is as nearly round as possible at the target, and that at no time does the storage surface reach saturation.

3.2 Resolution of Electrical-Visual Storage Tubes

A standard raster is written once and read visually under stated operating and viewing conditions. The individual lines of the written raster are readily discernible at large spacings but blend into a relatively uniform background as the spacing decreases. The tube is successively read, erased, and rewritten with the standard raster, but with lines spaced progressively closer together until the individual lines in the written pattern cannot be discerned visually. The resolution value measured at this condition is the *limiting resolution.*

The limiting resolution in lines per picture diameter is Nd/h where N is the number of lines in the raster, h is the height of the raster in the limiting-resolution condition, and d is the useful picture diameter or the diagonal.

The resolution specification should also include separate measurements of limiting resolution taken in orthogonal directions with identical tube-operating conditions.

To permit analysis of deflection defocusing and aberrations, a complete tube specification should also specify resolution under the following conditions:

a) The limiting resolution at the center with the beam focused at the center.
b) The limiting resolution along a diameter or diagonal, at a point 90 per cent of the distance from the center to the edge with the beam focused at the center.
c) The limiting resolution at the point specified in b) with the beam focused at that point.

Note: The values for the writing-beam current, raster size, scanning speed, and luminance of output should be included among the stated operating conditions.

3.3 Resolution of Beam-Indexed Electrical-Signal Storage Tubes

Under stated operating conditions, including the number of spots, a square array (in outline and detail) is inscribed within the useful storage surface. The beam is indexed once through this array in such a way as to write a checkerboard pattern: that is, adjacent elements are of opposite character (see Fig. 2). Writing-pulse length should be such that the spot under bombardment reaches 90 per cent of saturation in a single exposure to the beam. The beam is again indexed through the array and the reading process is carried out on all spots. With the spots in the array widely spaced, the output signals from both written and unwritten spots are said to be at their respective reference levels. The test sequence just described is repeated, increasing the density of spots per unit storage surface area (by reducing the size of the array) until the difference of the two types of readout signals shifts by a specified percentage of the difference between their respective reference levels. The resolution is defined in terms of the density of spots under this condition of degradation. The total number of such resolvable spots which may be written in the original inscribed area is now computed and given as a single number for the resolution of the tube under test.

Note: The surface must be completely erased before each writing period, either by the reading process or by a separate erasing cycle.

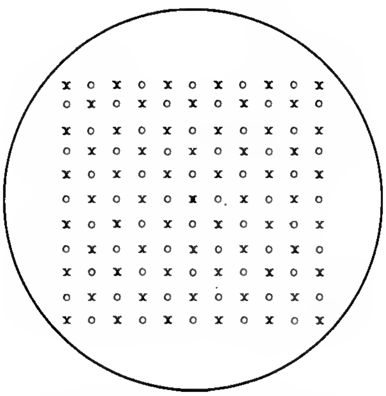

Fig. 2—Typical spot array for resolution test. X = test spot. O = unwritten spot.

4. Measurement of Writing Speed or Writing Time

The writing-speed capabilities of a tube is a function of such factors as the writing-beam current, the capacitance between the storage surface and other electrodes, and the secondary-emission ratio of the storage surface at the voltage used.

When the electron beam is indexed to discrete positions during writing, the concept of writing time per element is more applicable than writing speed and is measured as described in Sections 4.3 and 4.4.

4.1 Writing Speed of Scanned Electrical-Signal Storage Tubes

Under stated operating conditions, a standard raster is written once and read orthogonally. If the line spacing is sufficient to permit separated lines to be written, the peak-to-peak amplitude of the read signal is independent of writing speed at slow speeds, but decreases as the writing speed increases. Starting with a writing speed that produces saturation, the raster is successively erased and rewritten with higher write-scanning speeds at constant reading speed, until the signal amplitude is reduced to a specified fraction of the saturated value. A complete specification of writing speed is a curve of writing speed vs output level. Where a single value of writing speed is used, this should be the value for writing to 90 per cent of the saturation level.

4.2 Writing Speed of Scanned Electrical-Visual Storage Tubes

Writing speed may be measured simultaneously with the measurement of limiting resolution as described in Section 3.2. The conditions and procedures are preferably identical to those stated for that measurement.

The area rate of writing is hw/t, where h is the height of the raster for the condition of visually-limited contrast ratio, w is the width of the raster, and t is the time per raster.

The writing speed nw/t may be calculated by dividing the area rate of writing by the line width h/n, where n is the number of lines in a raster. The writing speed is completely specified by a curve of writing speed vs light output. Where a single value of writing speed is stated, this should be the value for writing to 90 per cent of the saturation level.

4.3 Writing Time of Beam-Indexed Electrical-Signal Storage Tubes

Under conditions and procedures identical to those stated for measurement of the resolution of beam-indexed storage tubes (Section 3.3), each test spot is written to saturation. The average amplitude of the signals from the test spots during the first read operation is noted. The storage surface is then erased and prepared for rewriting using a stated value of beam current. The time duration of the rectangular beam-intensifying pulse necessary to write a test spot to a charge level which is a specified percentage of the saturation level, typically 90 per cent of saturation, is determined. The duration of this beam-intensifying pulse shall be considered the writing time per element with the stated beam current (see Fig. 3).

4.4 Writing Time of Beam-Indexed Electrical-Visual Storage Tubes

Some electrical-visual storage tubes may be operated as beam-indexed tubes. In these types (*e.g.*, character-writing storage tubes) a writing-time test is used.

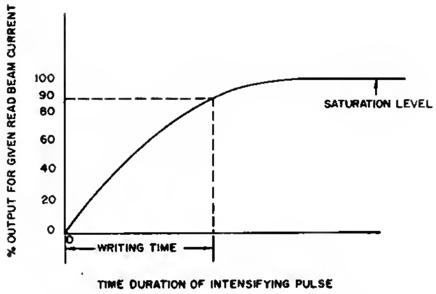

Fig. 3—Typical writing-time curve.

The storage surface is erased and, under stated operating conditions including a stated value of beam current, the time duration of the rectangular beam-intensifying pulse necessary to write a character, symbol, or bit is determined. The duration of this pulse is taken as the writing time per character, symbol, or bit with the stated beam current. For tubes using gray-scale storage, the percentage of saturation level to which the character is written must be specified.

Where a single value of writing time is stated, this should be the value for writing to 90 per cent of the saturation level. For tubes utilizing bistable storage, it is sufficient to say that a spot is written.

5. Measurement of Erasing Speed or Erasing Time

The erasing speed capabilities of a tube are a function of such factors as the magnitude of the erasing current to the storage surface, the capacitance between the storage surface and other electrodes, and the secondary-emission ratio of the storage surface at the voltage used.

When the electron beam is indexed to discrete positions during erasing, the concept of erasing time per element is more applicable than erasing speed, and erasing time is measured in the test described in Section 5.3.

In tubes using a flood beam to erase, a measurement of the time required to erase the complete storage surface is commonly used. This erase time test is also described below.

Note: Since priming is a special type of erasing, the test techniques used to determine priming speed are not specified. In practice the same tests are used as for erasing speed measurements, except that the tube operating voltages are set for priming during the test.

5.1 Erasing Speed of Scanned Electrical-Signal Storage Tubes

In some storage tube applications, area rate of erasure is often of equal significance to linear rate of erasure. For this reason, methods of measuring both parameters are described below.

To measure area rate of erasure, the erasing beam is scanned to form a standard raster, the size and location of which has been adjusted so that the raster falls within the boundaries of a writing raster established in accordance with Section 4.1. The writing conditions are adjusted so that a specified percentage of the saturation level is reached. It is further required that the lines forming the erasing raster overlap sufficiently to insure an essentially uniform erasing pattern. Under stated operating conditions, the storage surface is sequentially written, erased under the controlled conditions (test erased), read, and totally erased. Either the time per raster or the number of test-erasing rasters is increased until the test-erasing read signal from the area that has been test erased is reduced to a specified percentage of the saturation level. The area rate of erasure is hw/tN, where

h = height of test-erasing raster,
w = width of test-erasing raster,
t = time per test-erasing raster.
N = number of test-erasing rasters.

To measure linear rate of erasure or erasing speed, a standard raster is written in accordance with Section 4.1. The writing conditions are adjusted so that a specified percentage of the saturation level is reached. Reading for purposes of this test should be orthogonal to writing. The pattern is erased with lines well spaced. Writing, test erasing, reading, and total erasing are repeated sequentially. Either the number of test-erasing rasters or the time per raster is increased until the reading signal from the area that has been test erased is reduced to a specified percentage of the saturation level. The erasing speed is nw/tN, where n is the number of lines in the test-erasing raster, other terms having the significance quoted in the preceding paragraph. (For tubes that are erased by the reading process, a controlled reading raster performs the test erasing.)

5.2 Erasing Speed of Scanned Electrical-Visual Storage Tubes

In some storage tube applications area rate of erasure is often of equal significance to linear rate of erasure. For this reason methods of measuring both parameters are suggested below.

To measure area rate of erasure, the erasing beam is scanned to form a standard raster, the size and location of which have been adjusted so that the raster falls within the boundaries of a writing raster established in accordance with Section 3.2. The writing conditions are adjusted so that a specified percentage of saturation level is reached. It is further required that the lines forming the erasing raster overlap sufficiently to insure an essentially uniform erasing pattern. (In other words, no line structure is visible.) Under stated operating conditions, the storage surface is sequentially written, erased under the controlled conditions (test erased), viewed, and totally erased. Either the time per test-erasing raster or the number of test-erasing rasters is increased until the luminance of the area that has been test erased is reduced to a specified percentage of the saturation level. The area rate of erasure is hw/tN,

where

h = height of test-erasing raster,
w = width of test-erasing raster,
t = time per test-erasing raster,
N = number of test-erasing rasters.

The term *erasing speed* is applicable to only those electrical-visual tubes using a scanned erasing beam.

To measure linear rate of erasure or erasing speed, a standard raster is written in accordance with Section 3.2. The writing conditions are adjusted so that a specified percentage of saturation level is reached. The pattern is erased with lines well spaced. Writing, test erasing, viewing and total erasing are repeated sequentially. Either the number of test-erasing rasters or the time per raster is increased until the luminance of the display where it has been test erased is reduced to a specified percentage of the saturation level. The erasing speed is nw/tN, where n is the number of lines in the test-erasing raster, other terms having the significance quoted in the preceding paragraph.

5.3 Erasing Time of Beam-Indexed Electrical-Signal Storage Tubes

Using conditions and procedures identical to those stated for measurement of resolution of beam-indexed storage tubes (Section 3.3), each test spot is written to saturation. The average amplitude of the signals from the test spots during the first reading operation is noted. The storage surface is then rewritten and partially erased using the same beam current for the writing and erasing operations as was used previously to write to saturation. The average output amplitude during the subsequent read is determined. This operation is repeated using varying time durations of the erase beam-intensifying pulse. The time duration of the rectangular beam-intensifying pulse necessary to erase test spots to a level which is a specified percentage of the saturation level, typically 10 per cent of saturation, is taken as the erasing time per element with the stated erase-beam current.

5.4 Erasing Time for Flood-Gun Operation in Electrical-Visual Storage Tubes

The erasing time under stated erasing conditions is taken as the time required for the flood beam to reduce the output luminance from a stated initial value to a stated fraction of that value.

A rectangular raster as used in Section 3.2 is written to the specified luminance. The raster is then erased by pulsing the tube to erase conditions for a known length of time. The process is repeated, and the length of erase time is changed until the desired degree of erasure is obtained.

6. Measurement of Retention Time

The measurement of retention time in a charge storage tube is made without rewriting, reading, or erasing the storage element between the time the signal is written and the time the reading for measurement is taken. If holding action is used, the technique should be described.

Under stated operating conditions, a standard raster is written into storage to between 75 and 95 per cent of saturation. Line spacing must permit 100 per cent relative output amplitude (see Section 3.1). All electron beams of the tube (except possibly the holding beam) are then biased off for a measured time, and the voltages on other electrodes are set to suitable stand-by values. The output is read or viewed immediately upon completion of the measured time interval, and the percentage of the original output-signal amplitude is recorded.

Ambient temperature conditions during the test should be stated for tubes in which temperature of the dielectric markedly affects retention time.

7. Measurement of Reading Characteristics

In charge storage tubes, reading characteristics will depend upon the retention characteristics, effects of the reading beam upon the stored charge pattern, effects upon the stored charge pattern of ions formed by the reading beam, and effects of extraneous electrons resulting either from secondary or field emission.

7.1 Measurement of Read Number

The measurement of read number in a charge storage tube is made without rewriting. If holding action is used, the holding technique should be described.

Under stated operating conditions, a standard raster is written once to a percentage of saturation specified in the accompanying note and then read off repeatedly with a standard raster oriented orthogonally. From a plot of change in relative output amplitude near the center of the storage screen (in the region of ion-spot charge distortion) vs number of reading scans, the read number can be specified for any change of stored-pattern amplitude. The read number at the specified beam current, for a change in relative output amplitude to 50 per cent of its initial value, is recommended where a single value is used.

Note: For bilevel operation, greater than 90 per cent of saturation is specified; for gray scale operation, between 40 and 60 per cent is specified.

7.2 Read Time of Beam-Indexed Storage Tubes

Using conditions and procedures identical to those described for measurement of the read-around number of beam-indexed storage tubes (Section 7.3), each test spot is written to greater than 90 per cent of saturation. The average amplitude of the signals from the test spots during the first reading operation is noted. The storage surface is then read a second time and the change in value of the average amplitude of the signals is determined. The storage surface is erased and rewritten, and the above operation is repeated using varying time dura-

tions of the reading beam-intensifying pulse. The time duration of the beam-intensifying pulse which results in a prescribed percentage reduction in output signal between the first and second reading operations shall be considered the reading time per element with the stated reading beam current.

7.3 Read-Around Number of Beam-Indexed Storage Tubes

Under stated operating conditions, a square or rectangular array of spots covering the entire useful storage surface is established. Alternate rows and columns shall be composed only of interaction spots. All remaining spots, except at the pattern edges, represent test spots (see Fig. 4). Thus any test spot in the array is surrounded by eight interaction spots. An array of this type is written once. The test spots are read, the average output signal on the first reading operation being noted. A base-line clamp preceding each pulse may be used if desired; however, the specification should state whether or not it is used and at what level. The beam current and beam-intensifying pulse duration should be recorded. The test spots are rewritten with no current to the interaction spots. The reading operation is performed on the interaction spots a sufficient number of times to produce a specified change (20 per cent is recommended) in the average amplitude of the output from the test spots on the first read operation. The change is measured from the initial average amplitude to the final average amplitude, using a base-line clamp preceding each pulse if desired.

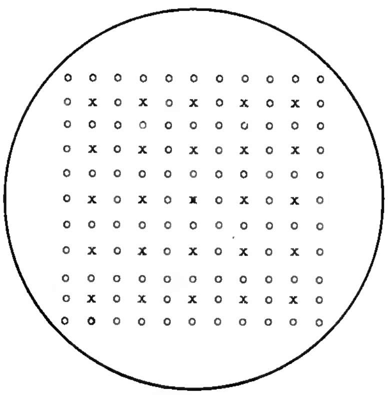

Fig. 4—Typical test array for measurement of read-around number. X = test spot. O = interaction spot.

Read-around number is specified as 6 times the number of reading frames used to produce the prescribed amplitude change.

Note: The number 6 is used since it has been assumed arbitrarily that each adjacent spot has an interference value of 1 and each corner spot of $\frac{1}{2}$.

The complete operation should be carried out in a period of time which is short as compared to the time in which the output will change by 20 per cent because of retention characteristics of the storage tube.

7.4 Measurement of Viewing Time for Visual Output Tubes

A standard raster is written to saturation luminance with overlapping lines. The luminance level near the center of the viewing screen is then measured. (For this measurement, the spectral response of the measuring device is unimportant, as only relative luminance readings are required.) The complete raster is then erased to a stated percentage of its initial luminance level, and the luminance is remeasured. The tube is read continuously and the time interval required for the luminance level to change (usually increase) by a stated percentage of the initial luminance level is measured. This is considered to be the viewing time for the value of reading-beam current used. Operating conditions, including any means utilized for extending the viewing time, must be specified.

Note: This viewing-time test is not intended for tubes designed to utilize bistable operation with continuous holding action. The viewing time based on changes in luminance level is essentially infinite for such types.

8. Measurement of Decay Time

There are two forms of decay in charge storage tubes. Static decay is caused by leakage along and through the dielectric at the storage element. Dynamic decay is the result of the action of the reading beam, ion currents, field-emission currents, and the holding beam (where used).

8.1 Static Decay Time

Static decay time may be measured in the same manner as reading characteristics (Section 7), except that all voltages appreciably affecting retention time are zero during each test period. (The static decay time commonly specified is the time interval during which the amplitude decays to 1/2.718 of its original value.)

8.2 Dynamic Decay Time

The measurement of dynamic decay time is identical with the measurement of reading characteristics (Section 7). The results are stated in terms of microseconds of reading per element for deterioration of relative output amplitude to a specified value.

The relation between read number and dynamic decay time is

$$\text{Dynamic Decay Time} = \frac{\text{Active Scan Time per Raster} \times \text{Read Number}}{\text{Number of Elements Contained in Raster Area}}.$$

The number of elements contained in the raster area is determined by resolution or read-around number and the portion of the storage area used.

9. Measurement of Signal-to-Shading Ratio

Shading is a gradual variation or a small number of gradual variations spatially fixed with reference to the

target area, as seen in the amplitude and/or background level of the output. Since there is usually no clear-cut delineation between shading and disturbance, the two must be separated on the basis of the number and/or type of variations. Gradual variations in the output of less than two or three per diameter or diagonal are termed shading. One or more abrupt variations or a high number of gradual variations is called disturbance. (See Section 10.)

9.1 Signal-to-Shading Ratio of Electrical-Signal Storage Tubes

For this measurement, the tube is operated with a scan along a single line only, commonly a diameter. Under stated operating conditions and with a single writing scan, several cycles of a fully resolved square wave, the base level of which corresponds to no signal storage, are written. The output produced by scanning the same line is examined, care being taken to insure that the writing and reading scans are exactly registered. The output signal will usually contain shading both of the background level and of the stored-signal amplitude. The ratio of the average output-signal amplitude A (see Fig. 5) to the peak-to-peak background shading B is the *signal-to-background-shading ratio.* The ratio of the average stored-signal amplitude A to the peak-to-peak shading variation of the stored signal $(S_{max} - S_{min})$ is the *stored signal-to-shading ratio.*

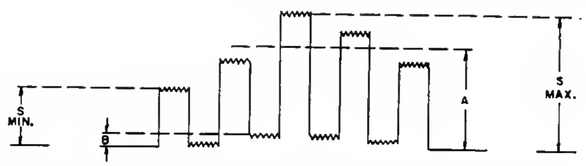

Fig. 5—Graphical representation of shading. A is the average output-signal amplitude. B is the peak-to-peak background shading.

9.2 Signal-to-Shading Ratio of Electrical-Visual Storage Tubes

With no signal stored, the viewing conditions are adjusted so that the luminance of the darkest portion of the tube is at visual extinction. By means of a photometer which measures light from a stated small area of the tube, the area of maximum background luminance is found and the luminance B (see Fig. 5) is measured. The entire surface is then written to such a level that some portion of the surface just reaches maximum luminance with no saturation at any point. With the same photometer, the area of maximum stored luminance is found and the luminance S_{max} is measured. The ratio of the average luminance A of the stored signal to the maximum background luminance B is the *signal-to-background-shading ratio.* The ratio of this average luminance A to the peak-to-peak variation $(S_{max} - S_{min})$ in the luminance of the stored signal is the *stored signal-to-shading ratio.* Localized blemishes are disregarded in making this measurement.

10. Measurement of Signal-to-Disturbance Ratio

Since a disturbance can have some structural detail near the limiting resolution of the tube, it is usually possible to increase the signal-to-disturbance ratio by operating the tube in such a manner as to decrease the resolution. Therefore, it should be measured in terms of a specified resolution.

When the disturbance has a periodicity (as from a mesh), the number of variations per tube diagonal or diameter should be specified.

10.1 Signal-to-Disturbance Ratio of Electrical-Signal Storage Tubes

The tube is operated under the same conditions as those used for measurement of resolution (Section 3.1), except that the single-line scanning and square-wave storage method described in Section 9.1 is used. The maximum level of the square-wave storage is written to just below saturation. During reading, the peak-to-peak variation on the top of the square waves C (see Fig. 6) is measured, as well as the peak-to-peak amplitude of the background disturbance D. Shading, isolated blemishes, and noise due to the output circuitry are disregarded in establishing the value of the ratio. The ratio of the average stored-signal amplitude A to the peak-to-peak background disturbance D is the *signal-to-background-disturbance ratio.* The ratio of this average stored signal amplitude A to the peak-to-peak disturbance variation of the stored signal C is the *stored signal-to-disturbance ratio.*

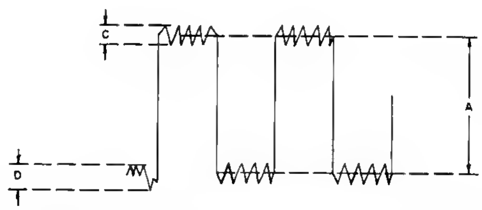

Fig. 6—Graphical representation of disturbance in electrical-electrical storage tubes. A is the average output signal amplitude. C is the peak-to-peak variation on the tops of the square waves. D is the peak-to-peak amplitude of the background disturbance.

10.2 Signal-to-Disturbance Ratio of Electrical-Visual Storage Tubes

The operating conditions are the same as those described in Section 3.2 wherever they affect the resolution. With no signal storage, the viewing conditions are adjusted so that the luminance of the darkest portion of the tube is just at extinction. Using a photometer having an effective aperture diameter equal to or smaller than the linewidth (refer to Section 9.2), the surface is examined along several diameters. The peak-to-peak variation D (see Fig. 7) in this background luminance is measured. The entire surface is then written to a level such that the maximum luminance is just below saturation. Again the surface is examined with the same photometer along the same diameters and the average lumi-

nance A of the stored signal and peak-to-peak variations C in this luminance are measured. Shading and localized blemishes are disregarded in establishing the above values. The ratio of the average luminance A to the peak-to-peak fluctuations in that luminance C is the *stored signal-to-disturbance ratio*. The ratio of the average luminance A to the peak-to-peak luminance fluctuations of the background D with no storage is the *signal-to background-disturbance ratio*.

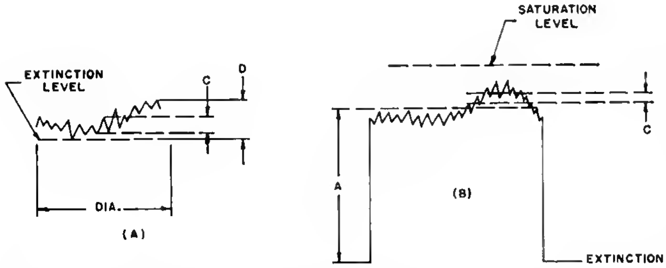

Fig. 7—Graphical representation of disturbance in electrical-visual storage tubes (A) with no storage, (B) with signal storage. A is the average output-signal amplitude at the high-brightness level. C is the peak-to-peak disturbance variation. D is the peak-to-peak amplitude of the background disturbance.

11. Luminance of Electrical-Visual Storage Tubes

11.1 Maximum Luminance

The maximum luminance S_{max} is the value determined in Section 9.2.

11.2 Contrast Ratio

The contrast ratio is identical to the *signal-to-background-shading ratio* A/B as measured in Section 9.2, for large areas, or the *signal-to-background-disturbance ratio* A/D as measured in Section 10.2, for small areas. (See Fig. 7.)

12. Measurement of Beam Current

The beam current in a cathode-ray storage tube is measured as the total current in an electron beam after the beam has passed through the final aperture of the electron gun.

13. Definitions

Abnormal Decay (Charge Storage Tubes). The *dynamic decay* of multiply-written, superimposed (integrated) signals whose total output amplitude changes at a rate distinctly different from that of an equivalent singly-written signal.

Note: Abnormal decay is usually very much slower than normal decay and is observed in bombardment-induced-conductivity type of tubes.

Barrier Grid (Charge Storage Tubes). A grid, close to or in contact with a *storage surface*, which grid establishes an *equilibrium voltage* for secondary-emission charging and serves to minimize *redistribution*.

Beam Current (Storage Tubes). The current emerging from the final aperture of the electron gun.

Bilevel Operation (Storage Tubes). Operation of a *storage tube* in such a way that the output is restricted to one or the other of two permissible *levels*.

Bistable Operation (Charge Storage Tubes). Operation of a *charge-storage tube* in such a way that each *storage element* is inherently *held* at either of two discrete equilibrium potentials.

Note: Ordinarily this is accomplished by electron bombardment.

Blemish (Storage Tubes). A localized imperfection of the storage assembly that produces a spurious output.

Bombardment-Induced Conductivity (Storage Tubes). An increase in the number of charge carriers in semiconductors or insulators caused by bombardment with ionizing particles.

Camera Storage Tube. A *storage tube* into which the information is introduced by means of electromagnetic radiation, usually light, and *read* at a later time as an electrical signal.

Cathode-Ray Charge Storage Tube. A *charge storage tube* in which the information is *written* by means of a cathode-ray beam.

Note: Dark-trace tubes and cathode-ray tubes with a long persistence are examples of cathode-ray storage tubes that are not charge storage tubes. Most television camera tubes are examples of charge storage tubes that are not cathode-ray storage tubes.

Cathode-Ray Storage Tube. A *storage tube* in which the information is *written* by means of a cathode-ray beam.

Charge Storage Tube. A *storage tube* in which the information is retained on the *storage surface* in the form of a pattern of electric charges.

Cloud Pulse (Charge Storage Tubes). The output resulting from space-charge effects produced by the turning on or off of an electron beam.

Collimate (Storage Tubes). To modify the paths of electrons in a *flooding* beam or of various rays of a scanning beam in order to cause them to become more nearly parallel as they approach the *storage assembly*.

Collimating Lens (Storage Tubes). An electron lens which *collimates* an electron beam.

Crossover Voltage, Secondary Emission (Charge Storage Tubes). The voltage of a secondary-emitting surface, with respect to cathode voltage, at which the secondary-emission ratio is unity. The crossovers are numbered in progression with increasing voltage. (See Fig. 8.)

Note: The qualifying phrase "secondary emission" is frequently dropped in general usage.

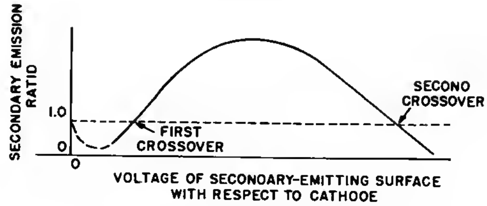

Fig. 8—Typical secondary-emission curve.

Cross-Talk, Electron Beam (Charge Storage Tubes). Any spurious output signal that arises from scanning or from the input of information.

Decay (Storage Tubes). A decrease in stored information by any cause other than *erasing* or *writing*.

Note: Decay may be caused by an increase, a decrease, or a spreading of stored charge.

Decay, Dynamic (Storage Tubes). *Decay* caused by an action such as that of the *reading* beam, ion currents, field emission, or *holding* beam.

Decay, Static (Charge Storage Tubes). *Decay* that is a function only of the target properties, such as lateral and transverse leakage.

Decay Time (Storage Tubes). The time interval during which the stored information *decays* to a stated fraction of its initial value.

Note: Information may not decay exponentially.

Destructive Reading (Charge Storage Tubes). *Reading* that partially or completely erases the information as it is being *read*.

Display Storage Tube. A *storage tube* into which the information is introduced as an electrical signal and *read* at a later time as a visible output.

Disturbance (Storage Tubes). That type of spurious signal, generated within a tube, which appears as abrupt variations in the amplitude of the output signal. These variations are spatially fixed with reference to the *target* area. (Note the distinction between this and *Shading*.)

Note: A blemish, a mesh pattern, and moiré present in the output are forms of disturbance. Random noise is not a form of disturbance.

Dynamic Range, Reading (Storage Tubes). The range of output *levels* that can be *read*, from *saturation level* to the level of the minimum discernible output signal.

Dynamic Range, Writing (Storage Tubes). The range if input *levels* that can be *written* under any stated condition of scanning, from the input that will write to saturation to the input that will write the minimum usable signal.

Note: The minimum usable signal is basically limited by the noise on the writing beam and by the transfer characteristics of the electron gun and the storage surface.

Electrical-Signal Storage Tube. A *storage tube* into which the information is introduced as an electrical signal and *read* at a later time as an electrical signal.

Equilibrium Voltage. See *Storage-Element Equilibrium Voltage*.

Erase (Storage Tubes). To reduce by a controlled operation the amount of stored information.

Erasing Rate (Storage Tubes). The time rate of *erasing* a *storage element*, line, or area from one specified *level* to another. (Note the distinction between this and *Erasing Speed*.)

Erasing, Selective (Storage Tubes). *Erasing* of selected *storage elements* without disturbing the information stored on other storage elements.

Erasing Speed (Storage Tubes). The lineal scanning rate of the beam across the *storage surface* in *erasing*. (Note the distinction between this and *Erasing Rate*.)

Erasing Time (Storage Tubes). The time during which the stored information is being *erased*.

Erasing Time, Minimum Usable (Storage Tubes). The time required to *erase* stored information from one specified *level* to another under stated conditions of operation and without rewriting.

Note: The qualifying adjectives "minimum usable" are frequently omitted in general usage when it is clear that the minimum usable erasing time is implied.

Escape Ratio (Charge Storage Tubes). The average number of secondary and reflected primary electrons leaving the vicinity of a *storage element* per primary electron entering that vicinity.

Note: The escape ratio is less than the secondary-emission ratio when, for example, some secondary electrons are returned to the secondary-emitting surface by a retarding field.

Flood (Charge Storage Tubes) (Verb). To direct a large-area flow of electrons, containing no spatially distributed information, toward a *storage assembly*.

Note: A large-area flow of electrons with spatially distributed information is used in image-converter tubes.

Half-Tones (Storage Tubes). Deprecated. See *Level*.

Hold (Charge Storage Tubes) (Verb). To maintain *storage elements* at an equilibrium voltage by electron bombardment.

Image Storage Tube. A *storage tube* into which the information is introduced by means of radiation, usually light, and *read* at a later time as a visible output.

Ion Charging (Charge Storage Tubes). *Dynamic decay* caused by ions striking the *storage surface*.

Ion Repeller (Charge Storage Tubes). An electrode that produces a potential barrier against ions.

Level (Storage Tubes). A distinct signal amplitude.

Levels, Usable (Storage Tubes). The output *levels*, each related to a different input, that can be distinguished from one another regardless of location on the *storage surface*.

Note: The number of usable levels is normally limited by shading and disturbance.

Memory Tube. Deprecated. See *Storage Tube*.

Moiré (Storage Tubes). A spurious pattern resulting from interference beats between two sets of periodic structures or scan patterns, or between a periodic structure and a scan pattern.

Note: In a *storage tube*, moiré may be produced by *writing* a resolved parallel-line scan pattern and *reading* using another parallel-line scan pattern. Since mesh elements usually have a periodic structure, moiré may also be produced between the mesh pattern and a scan pattern, or between two mesh patterns.

Nondestructive Reading (Charge Storage Tubes). *Reading* that does not *erase* the stored information.

Nonstorage Display (Display Storage Tubes). Display of nonstored information in the *storage tube* without appreciably affecting the stored information.

Overwriting (Charge Storage Tubes). *Writing* in excess of that which produces *write saturation.*

Prime (Charge Storage Tubes) (Verb). To charge *storage elements* to a potential suitable for *writing.*

Note: This is a form of *erasing.*

Priming Rate (Charge Storage Tubes). The time rate of priming a *storage element,* line, or area from one specified *level* to another. (Note the distinction between this and *Priming Speed.*)

Priming Speed (Charge Storage Tubes). The lineal scanning rate of the beam across the *storage surface* in *priming.* (Note the distinction between this and *Priming Rate.*)

Read (Storage Tubes). To generate an output corresponding to the stored information.

Read-Around Number (Storage Tubes). The number of times *reading* operations are performed on *storage elements* adjacent to any given storage element without more than a specified loss of information from that element.

Note: The sequence of operations (including priming, writing or erasing), and the storage elements on which the operations are performed, should be specified.

Read-Around Ratio (Storage Tubes). Deprecated. See *Read-Around Number.*

Read-Number (Storage Tubes). The number of times a *storage element,* line, or area is *read* without rewriting.

Read-Number, Maximum Usable (Storage Tubes). The number of times a *storage element,* line, or area can be *read* without rewriting before a specified degree of *decay* results.

Note: The qualifying adjectives "maximum usable" are frequently omitted in general usage when it is clear that the maximum usable read number is implied.

Reading Rate (Storage Tubes). The rate of *reading* successive *storage elements.*

Reading Speed (Storage Tubes). The lineal scanning rate of the beam across the *storage surface* in *reading.* (Note the distinction between this and *Reading Rate.*)

Reading Speed, Minimum Usable (Storage Tubes). The slowest scanning rate under stated operating conditions before a specified degree of *decay* occurs.

Note: The qualifying adjectives "minimum usable" are frequently omitted in general usage when it is clear that the minimum usable reading speed is implied.

Reading Time (Storage Tubes). The time during which stored information is being *read.*

Reading Time, Maximum Usable (Storage Tubes). The length of time a *storage element,* line, or area can be *read* before a specified degree of *decay* occurs.

Note 1: This time may be limited by static decay, dynamic decay, or a combination of the two.

Note 2: It is assumed that rewriting is not done.

Note 3: The qualifying adjectives "maximum usable" are frequently omitted in general usage when it is clear that the maximum usable reading time is implied.

Redistribution (Charge Storage Tubes). The alteration of the charge pattern on an area of a *storage surface* by secondary electrons from any other part of the storage surface.

Reflection Modulation (Charge Storage Tubes). A change in character of the reflected *reading* beam as a result of the electrostatic fields associated with the stored signal. A suitable system for collecting electrons is used to extract the information from the reflected beam.

Note: Typically the beam approaches the target closely at low velocity, and is then selectively reflected toward the collection system.

Regeneration (Storage Tubes). The replacing of stored information lost through *static decay* and *dynamic decay.*

Resolution (Storage Tubes). A measure of the quantity of information that may be *written* into and *read* out of a *storage tube.*

Note 1: Resolution can be specified in terms of number of bits, spots, lines, or cycles.

Note 2: Since the relative amplitude of the output may vary with the quantity of information, the true representation of the resolution of a tube is a curve of relative amplitude vs quantity.

Note 3: The resolution of storage tubes, when given in TV lines, refers to the number of black lines plus the number of white lines across the diameter of the storage surface. In television practice, resolution is measured vertically in a raster having a four-to-three aspect ratio.

Retention Time, Maximum (Storage Tubes). The maximum time between *writing* into a *storage tube* and obtaining an acceptable output by *reading.*

Saturation Level (Storage Tubes). The output level beyond which no further increase in output is produced by further *writing* (then called *write* saturation) or *reading* (then called *read* saturation).

Note: The word saturation is frequently used alone to denote saturation level.

Shading (Storage Tubes). The type of spurious signal, generated within a tube, that appears as a gradual variation or a small number of gradual variations in the amplitude of the output signal. These variations are spatially fixed with reference to the *target* area. (Note the distinction between this and *Disturbance.*)

Signal Integration. The summation of a succession of signals by *writing* them at the same location on the *storage surface.*

Storage Assembly (Storage Tubes). An assembly of electrodes (including meshes) which contains the *target* together with electrodes used for control of the stor-

age process, those that receive an output signal, and other members used for structural support.

Storage Element (Storage Tubes). An area of a *storage surface* that retains information distinguishable from that of adjacent areas.

Note: The storage element may be a portion of a continuous storage surface, or a discrete area such as a dielectric island.

Storage-Element Equilibrium Voltage (Charge Storage Tubes). A limiting voltage toward which a *storage element* charges under the action of primary electron bombardment and secondary emission. At equilibrium voltage the *escape ratio* is unity.

Note: Cathode equilibrium voltage, second-crossover equilibrium voltage, and *gradient-established equilibrium voltage* are typical examples.

Storage-Element Equilibrium Voltage, Cathode (Charge Storage Tubes). The *storage-element equilibrium voltage* near cathode voltage and below first-*crossover voltage.*

Storage-Element Equilibrium Voltage, Collector (Charge Storage Tubes). See *Storage-Element Equilibrium Voltage, Gradient Established.*

Storage-Element Equilibrium Voltage, Gradient Established (Charge Storage Tubes). The *Storage-element equilibrium voltage,* between first- and second-*cross-over voltages,* at which the *escape ratio* is unity.

Storage-Element Equilibrium Voltage, Second-Crossover (Charge Storage Tubes). The *storage-element equilibrium voltage* at the second-*crossover voltage.*

Storage Surface (Storage Tubes). The surface upon which information is stored.

Storage Time (Storage Tubes). Deprecated. See *Retention Time, Maximum* and *Decay Time.*

Storage Tube. An electron tube into which information can be introduced and *read* at a later time.

Note: The output may be an electrical signal and/or a visible image corresponding to the stored information. See also *Camera Storage Tube, Electrical Signal Storage Tube, Display Storage Tube, and Image Storage Tube.*

Target (Storage Tubes). The *storage surface* and its immediate supporting electrodes.

Transmission Modulation (Charge Storage Tubes). Amplitude modulation of the reading-beam current as it passes through apertures in the *storage surface,* the degree of modulation being controlled by the charge pattern stored on that surface.

Viewing Time (Storage Tubes). The time during which the *storage tube* is presenting a visible output corresponding to the stored information.

Viewing Time, Maximum Usable (Storage Tubes). The length of time during which the visible output of a storage tube can be viewed, without rewriting, before a specified *decay* occurs.

Note: The qualifying adjectives "maximum usable" are frequently omitted in general usage when it is clear that maximum usable viewing time is implied.

Write (Storage Tubes). To establish stored information corresponding to the input signal.

Writing Rate (Storage Tubes). The time rate of *writing* on a *storage element,* line, or area to change it from one specified *level* to another. (Note the distinction between this and *Writing Speed.*)

Writing Speed (Storage Tubes). Lineal scanning rate of the beam across the *storage surface* in *writing.* (Note the distinction between this and *Writing Rate.*)

Writing Speed, Maximum Usable (Storage Tubes). The maximum speed at which information can be *written* under stated conditions of operation.

Note: The qualifying adjectives "maximum usable" are frequently omitted in general usage when it is clear that the maximum usable writing speed is implied.

Writing Time (Storage Tubes). The time during which the stored information is being *written.*

Writing Time, Minimum Usable (Storage Tubes). The time required to *write* stored information from one specified *level* to another under stated conditions of operation.

Note: The qualifying adjectives "minimum usable" are frequently omitted in general usage when it is clear that the minimum usable writing time is implied.